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CORRECTION

On page 805 of The Proceedings for December, 1924 (Volume 12, Number 6), in footnote 3, line 4, change "Major Le Fry" to "Major Lefroy."

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A SUGGESTION FOR EXPERIMENTS ON APPARENT RADIO DIRECTION VARIATIONS*

By

L. W. Austin

(Communication from International Union of Scientific Radio Telegraphy)

The following facts have been observed in regard to the apparent direction shift of long-wave stations at moderate distances (200 to 700 km.).

(1) During the night, as is well known, these stations show large irregular shifts.

(2) During the day, from shortly after surnise, the bearing is generally nearly correct up to about two hours before sunset, tho on a few occasions, notably during some of the cold waves of January, 1924, there were shifts of a number of degrees early in the day.

(3) During the past year daily observations have been taken in Washington, on New Brunswick and Tuckerton, New Jersey, which show a regular shift toward the east beginning about two hours before sunset, reaching a maximum of from 8 to 15 degrees, returning to normal near sunset, after which there is frequently a sharp shift to the west followed by the irregular night shifts. This before-sunset shift is among the most regular phenomena so far observed in transmission in radio telegraphy as it has not failed once during the year. (There were no observations on Sunday and a few other days).

(4) The more distant stations, Rocky Point (415 km.) and Marion (660 km.) seem to show less regular sunset shift than those at about 250 km.

(5) Annapolis (54 km.) shows no definite sunset shift, but frequently shows a short-period (10-15 minutes) continuous shifting for hours at a time at any time of day amounting to three or four degrees.

These apparent direction shifts are generally believed to be connected with reflection from the boundary of a conducting

*Received by the Editor, October 25, 1924.

layer in the upper atmosphere. According to the hypothesis advanced by Eckersley ("Radio Review" II, page 60, 1921), the deviation is caused by a reflected wave which, coming down, strikes the radio compass coil without having previously passed along the surface of the ground. This ungrounded wave will in general not have its magnetic field horizontal, as in the case of a grounded wave, but the field will have a vertical component which cuts the top and bottom of the coil, thus producing an electromotive force which, to produce silence in the telephones, must be compensated by rotating the coil from the position of true minimum.

As Professor Kennelly has pointed out, experiments on direction shift form one of the most promising means of gaining information regarding the Kennelly-Heaviside layer. A determination of the distance at which the variations are a maximum will give an indication of the height of the layer, while observations on possible differences in the effect due to differences in direction, time of day, type of antenna, and so on, may give further useful information. All such information in regard to the upper atmosphere is of the highest importance in developing a complete theory of radio transmission. These observations can easily be made by anyone having a long wave receiving set and a four- to six-foot coil antenna, capable of rotation, with fifty or more turns. If the observer does not wish to work up his results for publication but prefers to send them to this laboratory, he will be given due credit when our work is published.

SUMMARY: The author suggests the systematic investigation of apparent shifts in direction of transmitting stations as indicated on a loop receiver. He describes briefly some of the effects so far discovered.

RECENT INVESTIGATIONS ON THE PROPAGATION OF ELECTROMAGNETIC WAVES*

By

M. Baeumler

(*Communication from the National Telegraph Engineering Bureau of Germany*) (*Telegraphentechnische Reichsamt, Berlin*)

Our knowledge of the causes of the modifications experienced by electromagnetic waves in their passage thru space has gaps in it. We have observed diurnal and annual variations of intensity of the waves, the magnitudes of which are different at different times, and for which no unimpeachable explanations have yet been found. We have also observed a diminution of field intensity with distance which is not in agreement with the theoretical law connecting these quantities. In addition, we have not yet been able to determine whether the daytime or the night-time value of the received electrical field is the normal value, that is, which of these values is to be regarded as the one which is in accordance with the theory. In order to clear up these questions, which are intimately connected with the propagation of electromagnetic waves, comprehensive measurements have been carried out at the National Telegraph Engineering Bureau of Germany (Telegraphentechnische Reichsamt).

I. Methods of Reception

An objective and quantitative measuring arrangement is required for the investigation of the propagation phenomena, in order that the variations of intensity at the receiving stations may be precisely followed. Quantitative measuring apparatus has been described, among others, by the following investigators: L. W. Austin,[1] J. L. Eckersley,[2] C. W. Pickard,[3] G. Vallauri,[4] M. Guierre,[5] H. J. Round and U. F. C. Lunnon,[6] R. Bown, C. A. Englund, and H. T. Friis,[7] and L. F. Fuller.[8] Some of these investigators have used the telephone as the indicating instrument. In spite of careful calibration of the receiving apparatus,

*Received by the Editor, July 21, 1924. Translated from the German by the Editor. This paper will also be published in German in 1, volume 2 of "Elektrischen Nachrichten-technik."

it is not possible to avoid entirely the disadvantages of the telephone, namely the insensitiveness of the ear and the fact that the readings are subjective and therefore, in part, dependent on the observer. The degree of uncertainty which may result from the use of the measuring arrangements depending on the ear is made abundantly clear by the work of A. Klages and O. Demmler[9] and of the writer.[10] In order to avoid these defects, an objective measuring method has been worked out at the National Telegraph Engineering Bureau by G. Anders, which method will soon be fully described.

The following is the basis of this method: The current produced in the receiving antenna by the distant transmitter, after suitable amplification, causes deflections of an electrometer proportional to the received current. The measurement of the antenna current is accomplished by means of an auxiliary transmitter current, which is adjusted accurately to equality with the original antenna current in amplitude and frequency, and which is measured by means of a barreter or a thermo-couple. By the utilization of the principle of the current transformer, it is possible accurately to measure considerably smaller currents than could be measured directly by the available instruments. The measurements can be carried out during the regular operation of the distant transmitting station because the thread element of the electrometer accurately follows the code signals of the station which is under observation. The subjective errors of the observer are avoided in our measuring method, as are also the physiological effects of atmospheric disturbances of reception which tend to spoil the note of the incoming signal and thus to make them seem weaker than they actually are. Our method therefore enables the completely objective measurement of received signals.

II. Mathematical Treatment of the Propagation Phenomena

F. Kiebitz, in his paper "On the Propagation Phenomena and Disturbances of Reception in Radio Telegraphy,"[11] has considered in detail the explanations which have been devised for the physical processes which occur in the propagation of electromagnetic waves. To study the spreading of the electromagnetic wave, we must know the value of the electric field which the distant transmitter produces at the receiving station. The following relations exist, according to Hertz and Barkhausen,[12] between the transmitter current of a continuous wave station, the received current, and the electric field at the receiving station:

$$F = \frac{120\,\pi\, I_1\, h_1}{\lambda\, d} \tag{1}$$

$$I_2 = F\,\frac{h_2}{r} \tag{2}$$

$$I_2 = \frac{120\,\pi\, I_1\, h_1\, h_2}{\lambda\, d\, r} \tag{3}$$

In the above equations, the symbols have the following significance (using practical engineering units):

F = electrical field at the receiving station in volts per meter
I_1 = transmitting current in amperes
I_2 = received current in amperes
h_1 = effective height of the transmitting antenna in meters
h_2 = effective height of the receiving antenna in meters
r = total resistance of the receiving antenna circuit in ohms
λ = wave length in meters
d = distance from the transmitter to the receiver in meters.

These relations are based on the assumption of a perfectly conducting ground in the form of a plane surface, and hold for wave propagation in a perfectly insulating medium without absorption, reflection, or refraction of the waves by the atmosphere. For wave transmission over a spherical surface, the above equations contain the additional term:

$$\sqrt{\frac{a}{\sin a}}$$

according to H. Poincaré,[13] J. W. Nicholson,[14], H. March,[15] and W. von Rybzinsky,[16] because the distances on a sphere are smaller than those in a plane in the ratio

$$\frac{\sin a}{a}$$

Here a is the angle subtended at the center of the sphere by the arc joining the transmitter and the receiver.

We thus obtain as the value of the electric field at the receiving station, assuming no losses due to absorption, reflection, refraction, or variable conductivity of the earth, the following:

$$F = \frac{120\,\pi\, I_1\, h_1}{\lambda\, d}\sqrt{\frac{a}{\sin a}}. \tag{4}$$

We shall refer to this value hereafter as the "theoretical value." It has been attempted to determine the losses in actual transmission empirically and theoretically. Empirical relations have been worked out by L. W. Austin,[1] L. F. Fuller,[8] and others. In addi-

tion to those previously mentioned, ([13-16]), A. Sommerfeld has worked out a theory of the propagation of electromagnetic waves taking account of actual losses.

In general, the effect of these losses is given by a term of the form

$$\varepsilon^{-\frac{Kd}{\lambda^M}}$$

in the equation of the received electric field, where ϵ is the base of the natural system of logarithms, d the distance, λ the wave length. K and M are numerical constants. The above term will be referred to hereafter as the "absorption factor," and takes account of all field reductions in the propagation of electromagnetic waves except those resulting from the theoretical diminution of amplitude with distance.

In practice, the value of the absorption factor found by Austin and Cohen[1] is most generally used. It was determined by daytime measurements on waves traveling oversea over distance up to 3,800 km. and has the value $\epsilon^{-\frac{0.0015\,d}{\sqrt{\lambda}}}$ wherein $K=0.0015$, $M=0.5$, and d and λ are expressed in kilometers.

L. F. Fuller[8] found the value $\epsilon^{-\frac{0.0045\,d}{\lambda^{1.4}}}$ for the absorption factor for measurements between San Francisco and Honolulu (3,800 km.) and Tuckerton to Honolulu (8,000 km.).

III. Experimental Results

(a) Simultaneous Measurements of Transmitted and Received Currents

With the measuring equipment described above, Division II (Research) of the National Telegraph Engineering Bureau in Berlin (Königgrätzstr. 20), and the Radio Research Station at Strelitz have been carrying out a series of measurements on received antenna currents to determine the field strengths of the signals from the American high power stations. The measurements were made on the signals from the stations at Rocky Point, Long Island (call letters WQK and WQL) and Marion, Massachusetts (call letters WSO), all of which are owned by the Radio Corporation of America. Dr. Alfred N. Goldsmith showed most keen interest in our investigations and willingly acceded to all our requests, and Dr. L. W. Austin has also frequently helped us in our work. We cannot refrain from expressing our heartiest thanks to these gentlemen at this time.

During July and August, 1922, the measurements were made by having the antenna currents of both WQK and WQL at Rocky Point recorded at 4 A. M., 9 A. M., and 3 P. M. Central European Time and simultaneously measuring the received currents at Berlin. The results are given in Figures 1 and 2. The

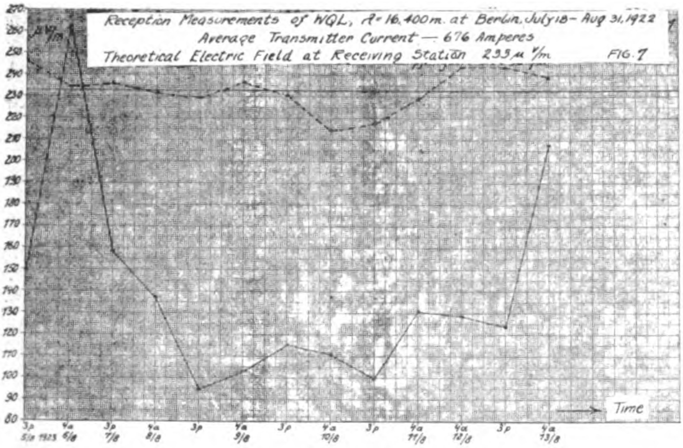

FIGURE 1

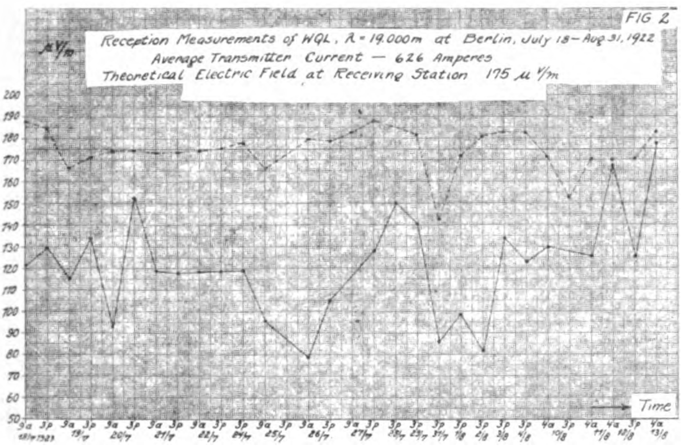

FIGURE 2

full line curve connects the points determined by the measurements, and the dashed line curve the points calculated by equation (4) from the actual transmitter currents; the latter curve is therefore directly dependent on the transmitter currents. It can be seen from the curves that the transmitter current varies only

slightly from its average value (at most 20 percent), whereas the field strength at Berlin varies as much as 90 percent. (For passing judgment on the relation between the propagation phenomena and the variations in received intensity, the variations in the transmitter current need not be here considered.) The numerical values of average transmitter currents, effective heights, distances, wave lengths, and resistances necessary for the calculations are given in the following table. The theoretical values of the electric field, on the basis of the average transmitting currents, are found from equation (4) to be 175 μv./m. for WQL and 233 μv,/m. for WQK. The actual values obtained by measurements, with the exception of a single night-time signal for WQL, are considerably less.

	Rocky Point			Marion
	WQK	WQL*	WQL	WSO
Wave Length........km.	16.4	19.0	17.5	11.6
Distance............km.	6,400	6,400	6,400	6,100
Average Transmitter Current............amp.	676	626	603	530
Effective Height of Transmitter Antenna.....m.	88	82.5	82.5	60.6
Effective Height of Receiver Antenna at Berlin.................m.	6.7*	6.7*	....	....
m.	5.6	5.6	5.6	5.6
Effective Height of Receiver Antenna at Strelitz................m.	19	19	19	19
Resistance of Antenna Receiving Circuit at Berlin.... ....ohms.	700*	1,220*	....	....
ohms	170	300	210	83
Resistance of Antenna Receiving Circuit at Strelitz..............ohms	462	....	790	194
Field Strength, Equation (1).............μv./m.	214	160	168	171
$\sqrt{\frac{a}{\sin a}}$	1.09	1.09	1.09	1.08
Field Strength $\times \sqrt{\frac{a}{\sin a}}$ μv/m.	233	175	183	185
Absorption Factor, Austin-Cohen................	0.0933	0.111	0.191	0.0685
Absorption Factor, Fuller.	0.568	0.626	0.59	0.414
Field Strength, Austin-Cohen..........μv./m.	21.7	19.4	16.8	12.7
Field Strength, Fuller μv./m.	133	....	....	76.5
Average Daytime Value μv./m.	150	....	....	60

*For the measurements during July and August, 1922.

(b) Simultaneous Measurements at Two Receiving Stations

No definite conclusions could be drawn from the first series of observations as to the relation between signal variations because there were considerable intervals between the taking of the measurements. It was, therefore, decided to diminish the intervals in question. Accordingly, the station at Strelitz, which was chosen as the second receiving station, was fitted up with apparatus of the same type as that used in Berlin. In November and December, 1922, simultaneous observations were taken at Berlin and Strelitz on the received currents from WQK. Figures 3 to 6 show the measured field intensities for several days. Sunrise and sunset are also shown, as well as the theoretical field strengths. It will be noted that there is a satisfactory agreement

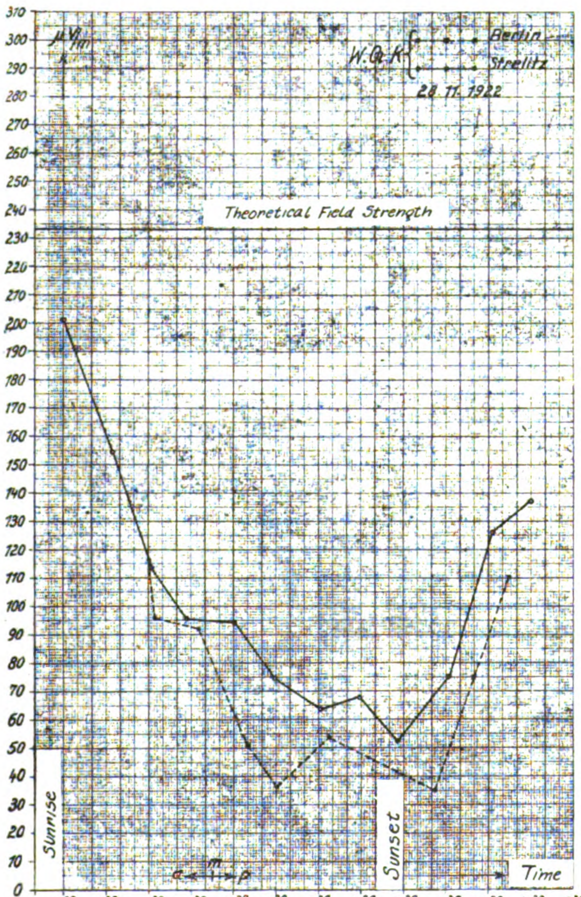

Figure 3

between the measurements in Berlin and Strelitz, particularly on December 6, 1922. The general shape of the curves indicates that the field strengths at sunrise are large, and indeed approximately equal to the theoretical value. With increasing elevation of the sun above the horizon, the signal strengths drop, reaching a minimum at noon or during the early afternoon hours, and gradually rising again after sunset. The general form of the daily changes in the electric field was the same for all days on which observations were taken and may, therefore, be taken as typica.. These effects are well known by those skilled in the handling of radio traffic. Loud signals (corresponding to large field strengths) are observed during the night, and weak signals (or small field strengths) during the day.

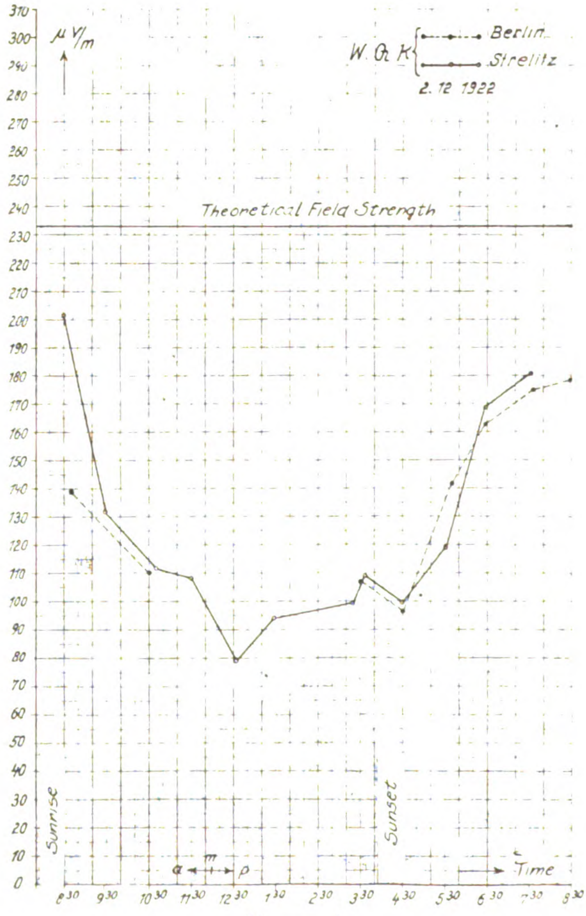

FIGURE 1

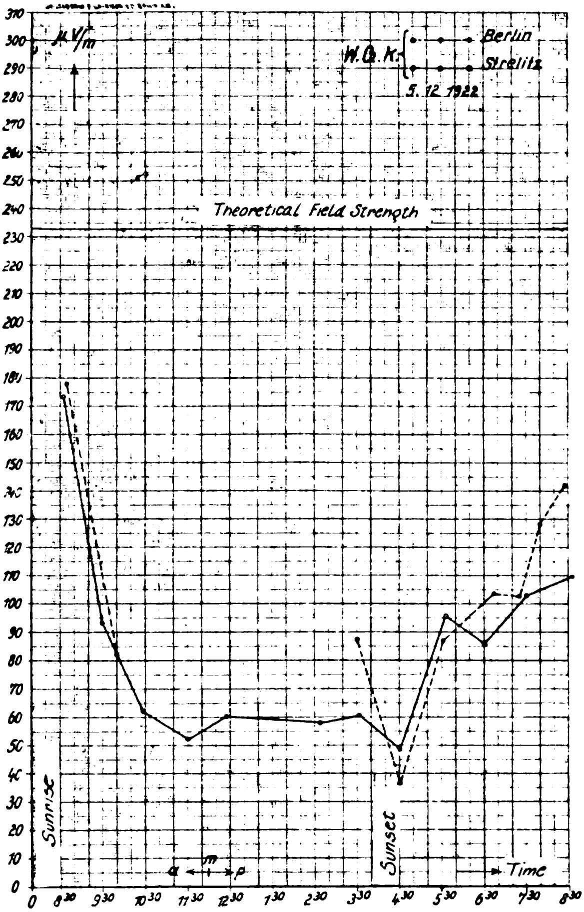

FIGURE 5

(c) CONTINUOUS OBSERVATIONS DURING SEVERAL DAYS AND NIGHTS

In order to get a still more complete picture of the propagation phenomena, particularly during the night, the scope of the observations at Berlin and Strelitz was broadened at the beginning of 1923. The received currents from WQK and WSO were observed for three successive days and nights each month. When WQK could not be heard, WQL was measured. Generally such observations were taken in the middle of the month, and at every forty minutes for each station. Individual measurements on the stations were therefore taken every twenty minutes. During these intervals, and particularly during the twilight and the

sunrise periods, many observations were made two or three minutes apart. The results of these measurements for the twelve months from February, 1923, to January, 1924, are given in Figures 7 to 18. The distribution of day and night between the transmitting and the receiving stations is shown in these figures. Sunrise and sunset at Berlin and Rocky Point are marked by short vertical lines on the axes of abscissas.

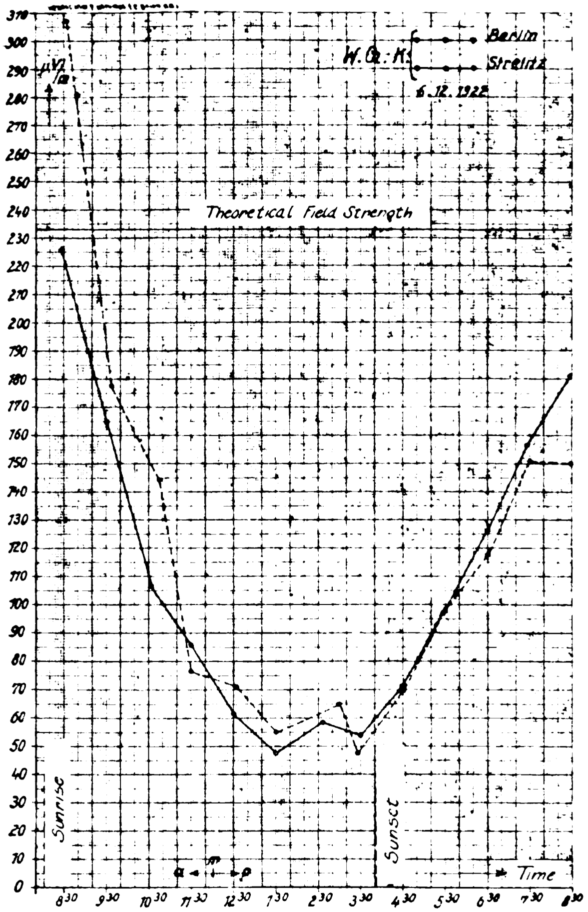

FIGURE 6

The thin horizontal lines indicate the time during which only one of the stations was illuminated by sunlight, and the heavy horizontal lines indicate the time during which the entire path of the waves lay in complete darkness. Times are given in Central European and American Eastern Standard Times. In addition,

there are shown on the curves the field intensities calculated according to equation (4) for Berlin and Strelitz on the basis of the average transmitter currents, and hence having the values

233 μv./m. for WQK
185 μv./m. for WSO, and
183 μv./m. for WQL.

The value for WQL being in practical agreement with that for WSO, it is not specially marked on the measurement curves of January, 1924. Inasmuch as the observation station at Strelitz had to be shut down at the end of December, 1923, all later observations could be taken only at Berlin.

(d) General Discussion of the Results of the Measurements

The four curves which appear in the Figures 7 to 18 have the same general shape and roughly maintain their relative amplitude ratios thruout. In general, the field strength of WSO is less than that of WQK, which is also true for the corresponding values of the field strengths calculated from equation (4). The superiority or greater consistency of the longer waves is evidenced by the closer approach to the theoretical values of the signals from WQK as compared to those from WSO. If we consider individual readings, it is found that the values for WQK and WSO frequently are different, altho there is good general agreement. These differences are to be ascribed to the limits of accuracy of the measuring equipment. Anyone who has watched radio telegraphic reception with a galvanometer or an electrometer is well aware that the moving element of the instrument is usually in rapid motion because of the incoming code signals and the atmospheric disturbances. This is particularly the case for received signals which are at the lower limit of measurement or are of the same order of magnitude as the atmospheric disturbances, which has frequently been the case during the measurements on the American high power stations. Readings of the electrometer in such cases were made only on dashes during which no strays occurred, and in such cases the deflection of the electrometer thread at the end of the dash was sharply defined. In addition to the above uncertainties in the readings, there were variations due to changes in the transmitter energy and in its frequency, interference from other transmitters, uncertainty in the value of the effective antenna heights and resistances and variations therein, and differences in the individual measuring equipments, all of which limit the accuracy of measurement. We cannot,

there we assume too extreme an accuracy for the result. We have taken for our results an accuracy of 30 percent, which embraces a range such that the differences of the Berlin and Strelitz signal measurements will be included within it in about 5,000 cases. Differences between these measurements up to 70 percent occur in only about 100 cases (or 2 percent of the whole), and for weak field strengths accompanied by powerful atmospheric disturbances. They thus constitute instances in which the observation or measurement limit had been reached.

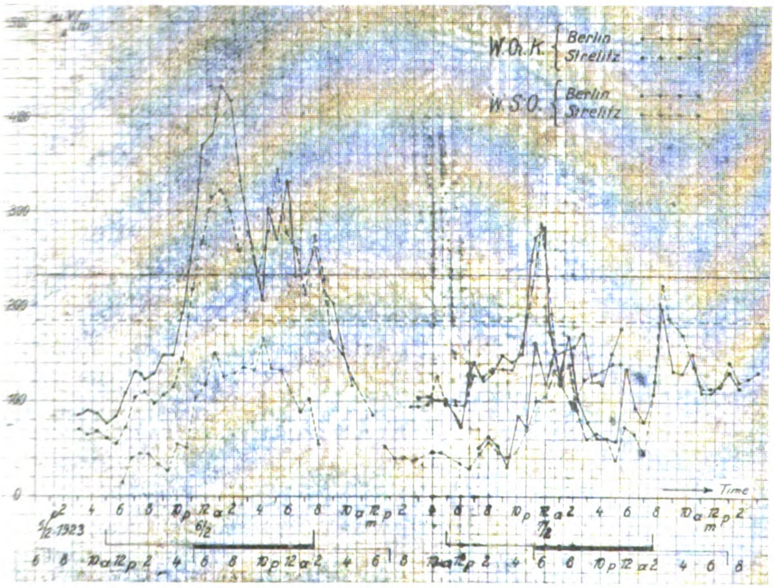

FIGURE 7

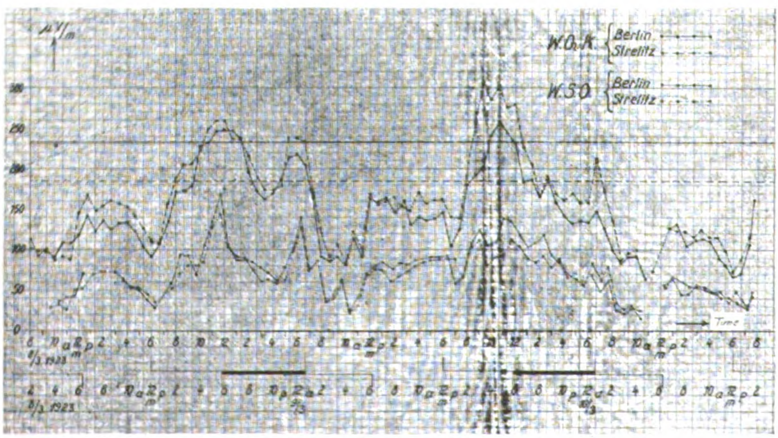

FIGURE 8

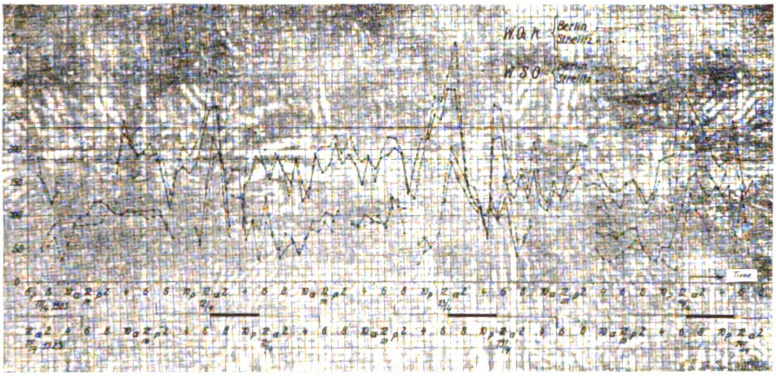

FIGURE 9

If we attempt to pass judgment on the accuracy of the measurements, we reach the conclusion that a probable accuracy of 30 percent in dealing with the very rapidly altering effects which must be followed by the measuring instruments represents a step forward in the field of radio telegraphic measurements. In connection with the measurement of the field strengths of the American high power stations, we carried out special tests to determine the reliability of our apparatus. Thru the courtesy of the Transradio Company for Overseas Radio Communication (Transradio A. G. für drahtlosen Übersee-Verkehr), arrangements were made to excite the large antennas at Nauen (at distances of 76 km. from Strelitz and 36 km. from Berlin) with such currents that the received currents at our observations were of the same order of magnitude as those produced by the American high power stations. These currents and the corresponding field strengths were in agreement with the theoretically calculated values, and larger currents also gave equally close agreement.

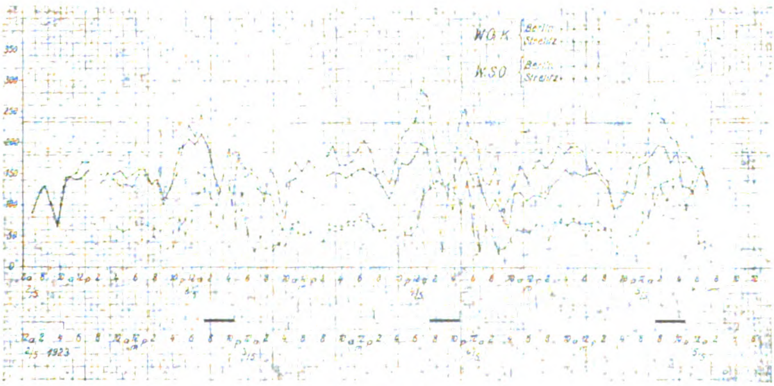

FIGURE 10

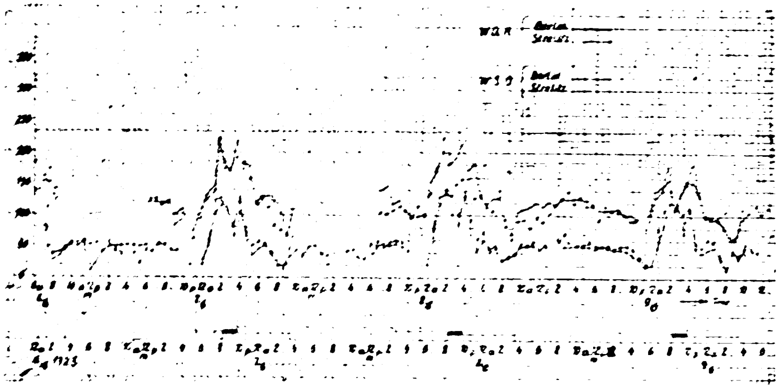

FIGURE 11

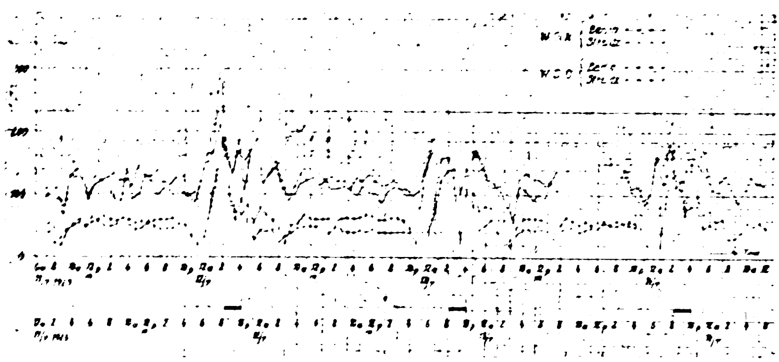

FIGURE 12

If we examine the curves more carefully with reference to diurnal and annual variations, we observe very clearly, as previously stated, the high values of the field strengths at night and the low values during the day. The increase in strength occurs after sunset at the receiving station, and the decrease at sunrise. The highest values are obtained when the entire span between the stations is in darkness. The night values are several times the day values, the ratio being approximately four-to-one in the winter and two-to-one in the summer. The differences, therefore, are not pronounced and are much less than the numerical values for these ratios previously given by other observers using, in some cases, subjective methods of measurement.

The differences between day and night values depend on the time of year. The night values are higher in the winter than in the summer. The least values have been measured at sunset at the receiving station and at sunrise at the transmitting station.

Between these two minima, there are found two maxima, namely, a marked one during the night and a relatively broad maximum during the day. During the winter the minima approach each other since sunset and sunrise occur relatively closely together at opposite terminals of the span between the stations during that season. The difference between the field strength by day in the summer and that in the winter is therefore chiefly a result of the approach of sunset and sunrise on paths between the transmitter and receiver of such lengths as we are here considering.

There are large and rapid variations in field intensity during the night in the winter and spring which are particularly remarkable. Instances of these occurred on February 6, 1923 at 2. A. M. and on May 4th at 1. A. M. The drop in signal strength, and

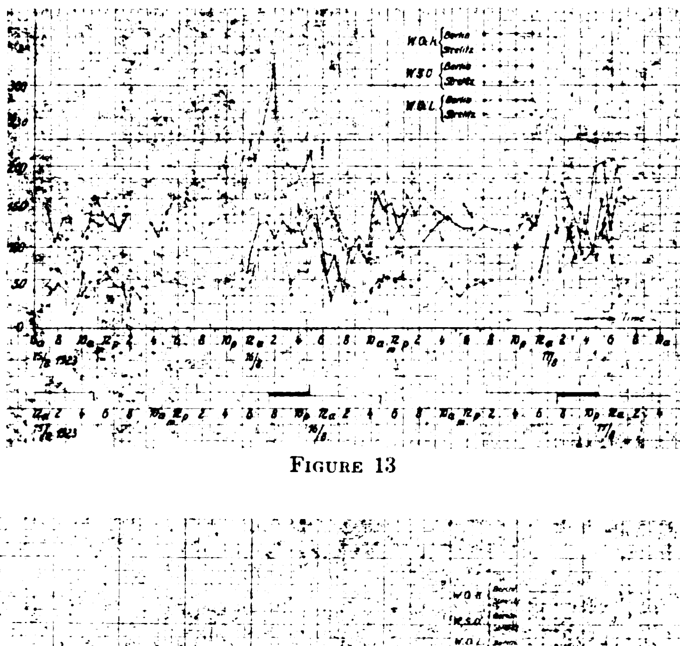

FIGURE 13

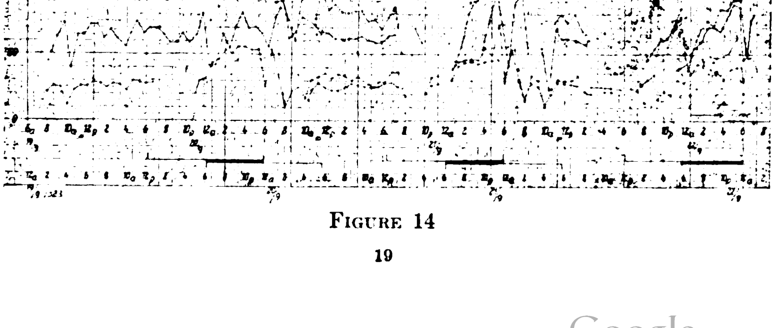

FIGURE 14

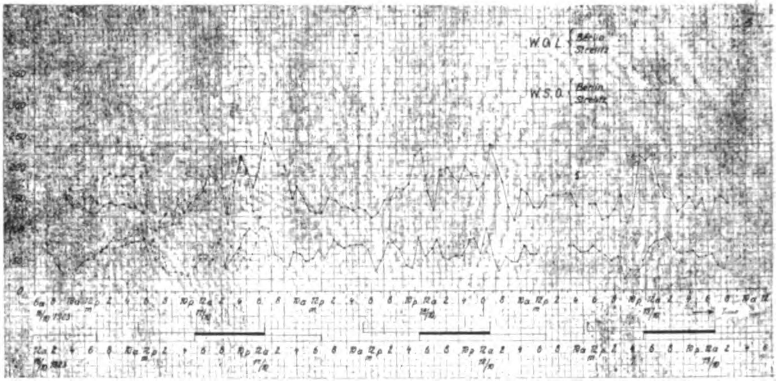

FIGURE 15

sometimes to a lower value than the weakest daylight signal, takes place in a short time, and is often accompanied, as we have noticed, by interruptions in the service. It first seemed correct to ascribe such signal changes to the transmitting station and due, for example, to a diminution in the transmitter current, alterations of frequency, or partial breakdown of the insulation of the transmitting antenna by rainfall. Questions addressed to Dr. Alfred N. Goldsmith showed these assumptions to be incorrect. The transmitter current was held almost perfectly constant during the measurement days, and it was easy to observe that at other times the transmitter currents and frequencies of the two stations were held extremely constant. No abnormal atmospheric conditions could be detected at the transmitting stations. Consequently the cause of the peculiar signal drop can only be found in the intervening medium.

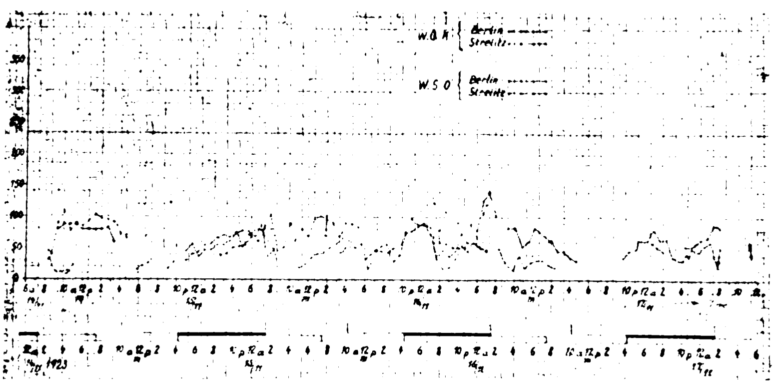

FIGURE 16

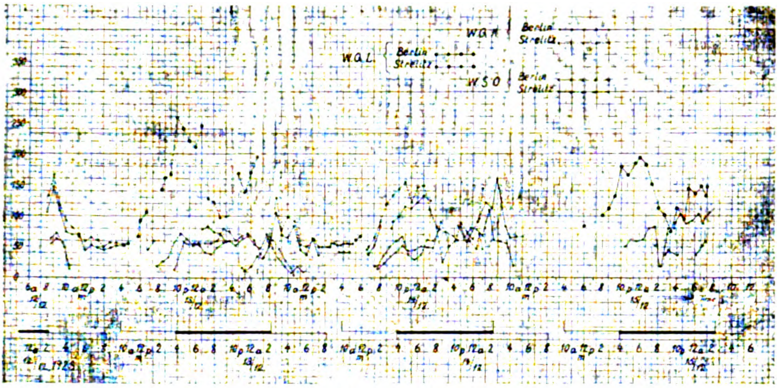

FIGURE 17

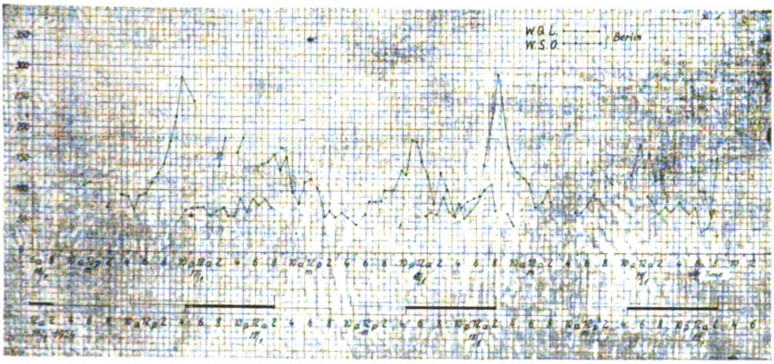

FIGURE 18

III. CONCLUSIONS

(a) THE NORMAL FIELD STRENGTH

In making assumptions as to the relationship between the day and night signals, one can either regard the day value as a diminution of the night value or the night value as an increase of the day value, and one can therefore be led to regard either the day value or the night value of the field strength as the normal one. The previous material enables us to give a decisive answer to the question of which value of the field amplitude is to be taken as normal. Using the law connecting field amplitude and distance, we should expect from equation (4) that, neglecting losses due to absorption, the following would be the electric field values at Berlin or Strelitz for the corresponding stations:

for WQK, an electric field of 233 μv./m., and
for WSO, an electric field of 185 μv./m.

If we compare these values with those actually measured, we find that they are exceeded only in a few instances in winter and spring during the night, and that the measured values during the summer lie below the theoretical values. The peak values at night during the winter and spring differ from the theoretical values by an amount less than the probable error of the measurements except for a few values in February, 1923 (for WQK), in August, 1923 (for WQL), and in January, 1924 (for WQL). Furthermore, the most recent measurements taken from February to June, 1924 (which are not given here), show that the night values of the measured field are of the same magnitude as the theoretical values.

In addition to the measurements of the American high power stations, the European high power stations at Lyons, Rome, Carnarvon, and Stavanger were measured at Strelitz. Stavanger shows a particularly marked difference between the day and night values—sometimes as much as eight-fold. A description of these observations will be given by S. Wiedenhoff[18]. If we carry out the same calculations for the Stavanger signals as we did in the case of the American stations, equation (4) gives a value of the electric field of 800 μv./m. Measurements at Strelitz gave actual night field strengths of from 200 to 1,100 μv./m., whereas the day value was only 130 μv./m. So that we find a good agreement between the theoretical value and the actual night field even for the comparatively short distance of 800 km. between Stavanger and Strelitz.

Since, as we have already seen, the field intensities calculated theoretically and the actually measured night field strengths are always comparable, we must regard the night values and the night transmission phenomena as normal, whereas the day values and the day transmission effects must be looked upon as abnormal or disturbed. This point of view has already been given by F. Kiebitz[19] in his paper on the refraction of electromagnetic waves in the atmosphere, and it is now confirmed by measurement.

Previously developed theories of transmission can be regarded as of possible validity only if they are based on assumptions similar to the above. For example, we might more readily regard the point of view of Fleming[20] as correct, which assumes energy losses during the day, than the theory of Eccles[21] in which the day transmission is regarded as normal and the night values are considered as due to an increase of field strength arising from a good conducting layer of ionized gas in the upper regions of the atmosphere—the so-called Heaviside layer. Our measurements,

by establishing such close agreement between the values of field strength calculated theoretically and the actually measured night values, make the assumption of the reflection of the waves by such a layer unnecessary.

(b) The Reason for Variation in Field Strength

As a result of the proof that we must regard the night field strength as normal, it follows that the day values are to be considered as a weakening of the night values. We therefore face the problem of the cause of this weakening and, more broadly, of field variations. As we have stated in III (d) of this paper, there is a unique correspondence between the daily changes in the field strength and successive positions of the sun and also the changes in illumination at both the transmitting and the receiving station influence the field strength. This typical daily variation in field strength has been noted in commercial service and also by other observers using subjective methods, the first of whom was G. Marconi. The material given in this paper verifies the earlier observations quantitatively by objective measurements, and has the great advantage that the actual magnitudes of the field strength variations are definitely determined, as previously described. It is therefore established that electromagnetic waves are enfeebled by day transmission. Whether this occurs directly or indirectly, and whether electrical or meteorological causes produce this effect is beyond our present knowledge.

Attention may, however, be directed to the following group of phenomena. The heating of the earth by the radiation of the sun brings about various stratifications of the atmosphere. At night, the air has been uniformly cooled and may be regarded as a homogenous layer. By day, the air becomes disturbed, and therefore, non-homogenous, by the irregular passage of heated air to the higher regions and the downward flow of the cooler air. Bodies of air of different densities lie side by side. The medium thru which the waves pass thus becomes "electrically turbid," and has a disturbing influence on the propagation of the waves since the waves are refracted, absorbed, or diffused at the boundary surfaces of air volumes of different densities, as was previously pointed out by F. Kiebitz.[19] After sunset the vertical motion of the air gradually ceases. The atmosphere slowly becomes uniform, and the causes of the disturbance of a perfectly free wave transmission are decreased. So that we may regard the night atmosphere as a meteorologically homogenous medium which interferes little or not at all with the passage of the waves.

The assumption that the waves at night pass thru a medium which does not absorb, diffuse, or reflect them is quite plausible and is supported by our measurement establishing the agreement between the theoretical and the actual field strengths.

In accordance with optical laws, the waves can be refracted only by masses of air of different densities if these masses have sufficiently large dimensions. Consequently field strength variations must be more common on the shorter waves because the formation of small masses of rarefied or compressed air can occur more readily and therefore more frequently than the formation of larger masses. As a matter of fact, more rapid and extreme variations of signal strength are observed on the short waves than on the long waves.

In part III (d) of this paper attention was directed to the fact that the measured night values of the field strength were greater in winter than in summer. This annual variation of the field strength fits readily into the above theory of the causes of field strength variations. During the winter the difference between the night and day values of ground and air temperature is small and therefore the "turbidity" of the atmosphere is small, so that a good homogenous intervening medium between the transmitting and the receiving station can readily be formed at night. In summertime, because of the marked heating of the ground and the short duration of darkness, the formation of such a homogenous atmosphere can occur only to a small extent, and thus the atmosphere remains non-homogenous and weakens the waves in their path even at night.

In addition to these causes of variable transmission, we have changes in the pressure of the atmosphere, the formation of clouds, precipitation of moisture, and accompanying phenomena. It was not possible to determine the extent to which these various meteorelogical conditions affect variations of field strength since comprehensive data relative thereto, particularly over the Atlantic Ocean, could not be obtained. The marked variations of the night field strengths occurring during the winter months cannot be explained by the above theory; their explanation must, therefore, be sought elsewhere.

(c) Field Strengths in Large Cities and in Flat Open Country

We are often told that the field strength produced by a station in a large city is less than that in open country because the layer of fog or smoke over the city weakens or diverts the waves.

If we consider our observations, we find this statement is not confirmed for the observed wave lengths of 16,400 and 11,600 m. Considering WQK and WSO, we find that the field strengths are sometimes higher in Berlin and sometimes in Strelitz, as can be seen directly from the curves.

(d) The Absorption Factor

The difference between the theoretical and measured values of field strengths is described as due to the sum total of all losses caused by absorption. Even a brief examination of the experimental curves shows the extraordinary difficulty which is necessarily experienced in attempting to express them by any equation. Such an equation would have to contain the distance, the wave length, the time of day and year, the weather, and even the nature of the intervening ground (because of the different velocities of electromagnetic waves over land and sea). It requires no further consideration to see that it is impossible at this time to produce such a universally applicable formula. The problem of the propagation of electromagnetic waves requires much further clarification thru numerous objective measurements before we can expect to derive such a formula.

If we do not agree to abandon entirely the development of an equation of this sort, we are forced to restrict it to some special case, for example for the transmission of the waves oversea by day. We shall examine the Austin-Cohen and Fuller formulas, using these limitations.

Applying it to our transmission measurements, the Austin-Cohen formula gives the following electric field strength values:

for WQK, a field of 21.7 μv./m. and
for WSO, a field of 12.7 μv./m.

Our measured field strengths are several times greater than these values. The daytime field strengths as measured for WQK are six times the value given by the Austin-Cohen formula, and for WSO they are five times the value. On the other hand, the absorption factor given by L. F. Fuller gives values in better agreement with our measurements, namely the following electric field strengths:

for WQK, a field of 133 μv./m. and
for WSO, a field of 76 μv./m.,

which are of the same magnitude as the actual measured values.

In measuring the field strength of Annapolis in the neighborhood of Rome, Vallauri found values which were thirteen times as large as those calculated, using the Austin-Cohen absorption

factor. Dr. Austin, who is carrying on regular measurements on the European high power stations at Nauen and Lafayette (Bordeaux), found field strengths during the day which were about twice as great as values calculated, using his absorption factor. Similar conclusions have been reached by G. W. Pickard, J. L. Eckersley, and C. R. Englund.[22]

We have seen that the results of measurements are to some extent in disagreement. The explanation for these differences remains to be found. We shall accordingly carry forward our measurement work.

SUMMARY: Since the summer of 1922, quantitative measurements have been carried out on the signal strengths of the American high power stations WQK and WSO, using an objective measuring method. The object of these measurements is to study the propagation of electromagnetic waves. It has been shown that these phenomena can be studied only by continued measurements.

Curves are presented showing the field strengths on three successive days and nights once each month for the year February, 1923-January, 1924. In view of the agreement of the calculated values of field strength and the values actually found at night, it is concluded that the night value is to be regarded as the normal one and the day value as the abnormal or disturbed one. An explanation of the diminution of field intensity is given by assuming that the atmosphere is "electrically turbid" by day in consequence of the heating of the earth and the resulting vertical motion of masses of heated air. The waves are refracted, absorbed or reflected (and hence weakened) at the boundary surfaces of air masses of different densities. Diurnal and annual variations of field intensity can be readily explained by this theory. It was not possible to establish any difference between field intensity in a large city and in nearby open country.

The derivation of a universally applicable formula giving the field strength, while taking account of all absorption losses, is regarded as impossible at present. The empirically determined absorption factor of the Austin-Cohen formula does not give results in agreement with measurement, but yields markedly smaller values of field intensity; on the other hand, the absorption factor found by L. F. Fuller gives values in good agreement with the results of measurement.

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THE MARCONI MARINE RADIO DIRECTION FINDER*

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Introduction

Radio direction finding is daily becoming an essential feature of mobile radio stations and today there exist a large number of steamers fitted with this extremely useful adjunct to navigation.

The introduction of the thermionic tube to the radio art may be said to have been mainly responsible for the practicability of the modern direction finder, as hitherto the range of these instruments was extremely limited, owing to the poor receptive quality of the loops necessarily employed when used in conjunction with the detectors available in pre-war days.

The erection of a large number of radio fog stations has given a large impetus to the fitting of direction finders on steamers, and the United States of America have shown great partiality towards these beacon stations, a large number of these radio beacons having been installed on lighthouses and light vessels by the Bureau of Lighthouses.

Historical

It is not the intention of this paper to go deeply into the theoretical side of the directional properties of antennas, but rather to describe the practical side of the modern marine radio direction finder. Bearing this in mind it will perhaps not be out of place to briefly trace the modern day apparatus from the early day experiments of radio directional effects.

Marconi in his early investigation noticed a marked directional effect associated with certain types of antennas and adopted such forms of antennas in his first trans-Atlantic work.

Prior to this, Hertz himself, had shown how electromagnetic waves of small length could be directed by reflection using parabolic mirrors. This fact is mentioned merely "*en passant*" in

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view of the interest which is being shown today with the radio beams, which, however, employ a different method to obtain directional reception than is to be described herein.

The well known inverted "L" antenna, such as is employed largely on board ship, has a very marked directional effect towards the heel of the "L," and in 1906 Marconi took out a patent embodying a series of these inverted "L" antennas, erected with the horizontal limbs equally spaced radially about the actual receiving apparatus.

However, progress in direction finding was given a practical start when certain investigators examined the properties of two spaced antennas, and of the receptive qualities of loops.

Time does not permit in this historical survey to discuss all the various types of apparatus, but it is perhaps necessary to mention that Messrs. Brown and Stone discovered that two open vertical antennas joined together and spaced half a wave length apart had marked directional effects, and it was proposed to rotate this system as a practical direction finder.

A scheme for overcoming the difficulty of rotating such an antenna system was introduced in 1907 by Bellini-Tosi, and their method forms the basis of the direction finders now utilized by the British Marconi Companies.

This system has many advantages over the rotating loop for the direction finder, as a loop, particularly of large dimensions, does not offer a very good mechanical proposition as regards rotation, whilst a small loop introduces electrical errors.

The Bellini-Tosi Direction Finder

Description of Basic System

The Bellini-Tosi Direction Finder consists substantially of three essential parts namely—

1—The special oriented antenna system.	Active portion of system.
2—The radiogoniometer.	Screened portion of system.
3—The amplifying and detector unit.	

The original method consisted in erecting two directional loops or frame antennas of equal dimensions insulated from each other and at right angles. The leading-in wires from these were brought thru some form of tuned buzzer device and connected to the radiogoniometer.

The radiogoniometer consisted of two antenna coils wound on a common frame and which were of equal electrical dimensions, wound at right angles to each other. Within this former,

carrying these two antenna coils, a rotatable coil, known as the search coil, was fitted.

The actual antennas and the respective coils of the radiogoniometer were connected in series with a variable condenser to allow of each antenna system being carefully tuned to the wave length of the station it was desired to receive.

One of these condensers had connected across it a small variable "billi" or tubular condenser to allow of fine tuning in order to produce perfect balance between the two antennas. The search coil was also shunted by means of a further condenser to permit of that circuit also being tuned to the incoming waves, and this search coil circuit was then connected to some form of tube receiver and amplifier.

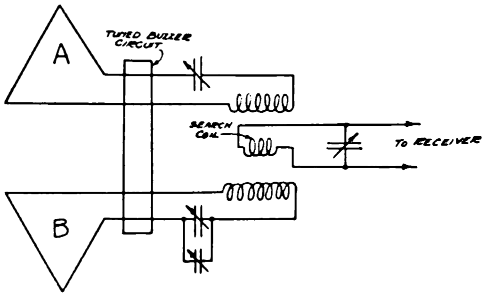

FIGURE 1

METHOD OF TUNING SYSTEM

The tuning buzzer was set so as to produce a wave corresponding to that it was desired to receive. Then having first tuned the search coil circuit and receiver to that wave, each loop was tuned separately without the second loop being in circuit.

Following on, the two loop circuits were then brought into action with the tuning buzzer still operating, and the radiogoniometer search coil was varied until a position was arrived at where the two antenna radiogoniometer coils produced no energy in the search coil and consequently no signals were heard in the receiver. This state of balance was not always effected immediately, and it was sometimes necessary to vary the final adjustment condenser shown across the tuning condenser of antenna *B*.

Altho it will be seen there are quite a large number of tuning adjustments to be made irrespective of the actual detecting and amplifying device, it was found that telegraphists readily became capable of accurately tuning the whole system very rapidly, within a short specified time. However, the method of tuning

this whole system is somewhat laborious and is not practical for everyday marine direction finding work.

The description of this early apparatus forms, however, a useful basis to work on in the developing of the modern day apparatus.

THEORY OF THE SYSTEM

Assuming the circuits all to be in resonance and regarding merely the oriented antenna system and radiogoniometer, now consider the effect of an electromagnetic wave incident on this system. (See Figure 2.)

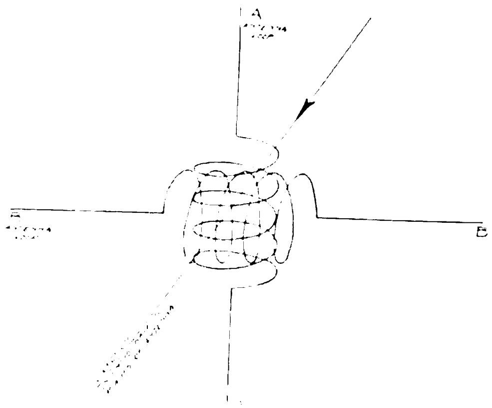

FIGURE 2

Suppose the wave is arriving in the plane of the antenna loop A, then following the usual electrical laws, no current will be induced in the antenna loop B which is at 90° to A, and consequently at 90° to the direction of the path of incoming waves.

But antenna A is so disposed as to be able to obtain the maximum effect of the incoming wave and will have a current induced in it producing a magnetic field along the cylindrical axis of the A antenna coil. The search coil rotating within the two antenna coils will have to be rotated to such a position as to be under the maximum coupling effect in order to produce a maximum current in its winding.

Similarly, if the incoming wave is exactly incident with the plane of the antenna "B," a magnetic field will only be produced along the axis of the antenna coil "B."

Now take the case of a wave arriving in a direction somewhere between the planes of the two antennas, then both antennas and their respective coils will have currents induced in them, the

relative strengths of each being dependent upon which of the antennas most closely coincides in direction with that of the incident wave.

It will be easily seen from this that the resultant field produced by the antenna coils bears exactly the same space relationship to the axis of these antenna coils as the direction of the received wave does to the planes of the antenna loops.

In order to obtain a maximum current in the search coil, it follows that the same must be rotated within the antenna coils until the cylindrical axis of the winding of this coil lies along the direction of the resultant electrical field produced by the two antenna coils. It follows, therefore, that at a position of 90° to this maximum position, a minimum or zero current point is obtained.

It is usual for the zero position of the search coil to be used for practical working as it is easier for the human ear to determine the minimum signal strength than to observe a maximum.

As the search coil is rotated thru 360°, it will be found that the signals heard in the telephones alternate maximum and minimum signals 90° apart;—in other words two maximum and two minimum points are given at 180° apart.

A pointer is fixed to this search coil which moves round a scale fixed to the radiogoniometer marked off in degrees, so that the direction of the resultant field can be obtained, and therefore the direction of the incident wave relative to the antennas is shown.

In ship work, of course, the relative directions of the antennas to the incident wave are continually varying and consequently the course of the ship has to be noted at the time when readings on the system are being observed.

Description of the Modern-Day Marine Direction Finder

It has already been pointed out that the tuned direction finder system offers rather a laborious manipulation if rapid and reliable working is required, and consequently a very much simplified apparatus embodying the direction finder and amplifier has been evolved (Figure 3).

Two antenna loops are erected and insulated from each other, one of which is rigged accurately fore and aft along the centre line of the vessel, and is known as the fore and aft loop, whilst the second is erected thwartships and is termed the thwartship loop.

These loops should be constructed to be as large as available space permits, having due regard to the local conditions and

should be erected as far as possible from ventilators, skylights, and other metal obstructions.

It is very essential that no portion of the loops come nearer than 6 feet to such obstructions, otherwise their presence will become known by the introduction of errors. Earthed metal constructions, if not symmetrically disposed under the loops, are also likely to cause errors if within twelve feet.

It has been noticed that the turning of a ventilator within the 12-foot limit has produced a very noticeable error varying with the position of the aperture of the ventilator.

The thwartship loop as a rule runs nearly the full beam of the ship, and less than 40 feet in this connection is not regarded as satisfactory. The fore and aft loop should be constructed so that

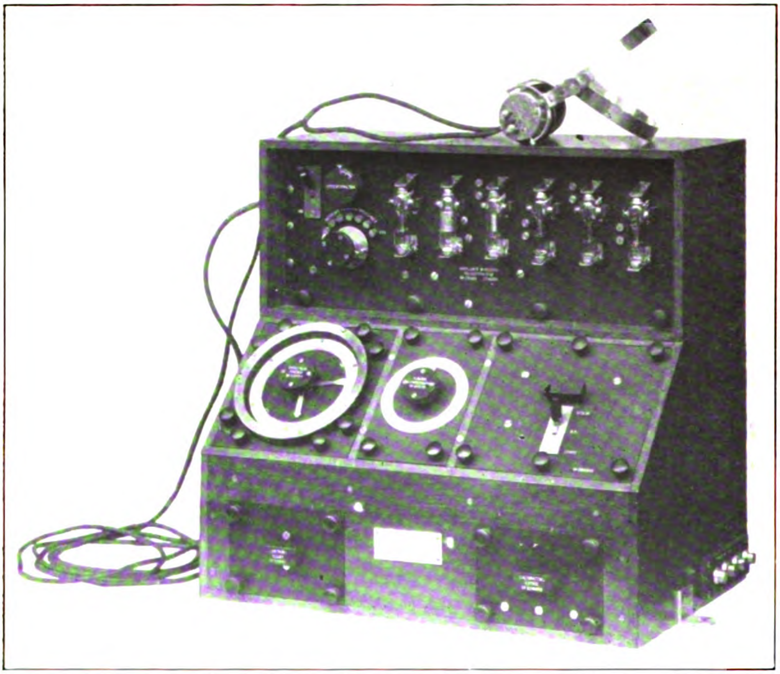

FIGURE 3—The Marconi Marine Direction Finder Type 11-B

its area is about three-quarters that of the thwartship loop, for reasons which will be discussed later. It is not necessary for both loops to be identical with regard to shape, neither is it essential that they should intersect each other altho this latter condition is generally desirable when viewed from a practical erection point of view.

At the time of erection the fore and aft loop is generally not made a permanent fixture as it is with this antenna that calibration alterations are made. Some typical types of direction finder antenna systems for shipboard use are shown in Figures 4 and 5.

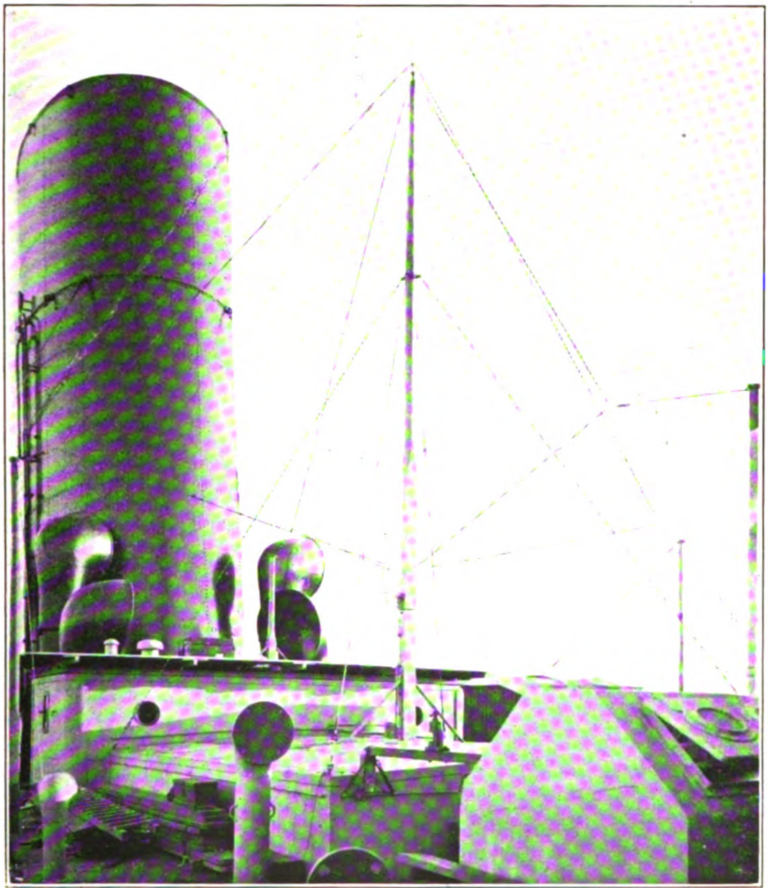

FIGURE 4

As a general rule, the leads from these loops are taken to four swan-neck insulators which are erected on a wooden spar fitted directly under the apex of the whole antenna. The leads from these insulators are taken to an outside junction box, each loop having its own box. These general external connection arrangements can be seen in the illustrations referred to above.

From these outside junction boxes the antennas or active system of the direction finder are connected by means of a special

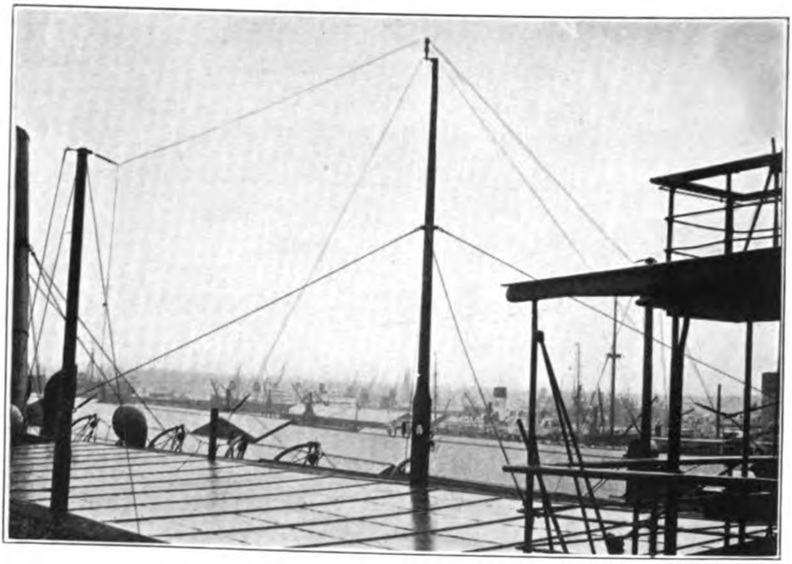

FIGURE 5

pair of cables to two further junction boxes located inside the radio room from which direct connection is made to the actual radiogoniometer (Figure 6). The cable effecting this junction

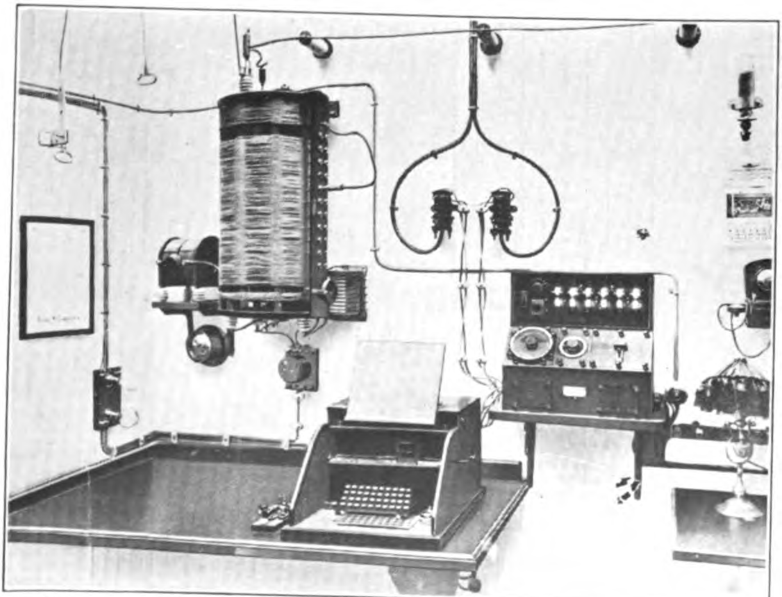

FIGURE 6—Connections of Radiogonimeter. (Inside Wireless Room).

is of a special design and is lead-covered and contains two wires insulated from each other by means of paper insulation which reduces the electrical capacity between the two contained wires of the cable to an absolute minimum. The actual capacity of this cable between the wires for every 100 feet is only 0.0008 mfds.

The reason for employing such a cable is to insure that the natural wave length of the complete antenna circuits is always shorter than the shortest wave it is desired that the radiogoniometer will have to receive.

In order to reduce the actual tuning adjustments to a minimum, these shipboard antennas are used in conjunction with a very tightly coupled radiogoniometer (Figure 7).

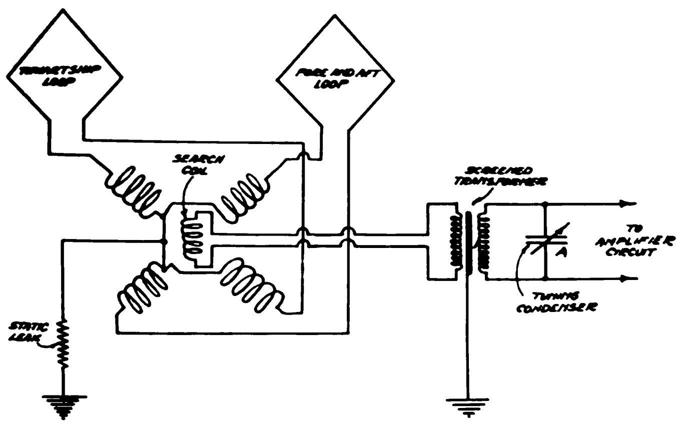

FIGURE 7

This action permits of the two loops being of an untuned or practically aperiodic order which are tunable by one condenser A, located in the coupled search coil. The actual coupling between this circuit and the two loops is somewhere of the order of 80 percent.

The goniometer itself consists of the two usual identical antenna coils disposed at right angles, the mid-points of which are earthed thru a static leak to insure of the antennas being protected from static discharges.

The introduction of a long cable connection between the active and screened portions of the direction finding system, such as is mentioned above, is likely to produce an undesirable electrical component which would be exaggerated if the loops were grounded directly at the mid-points of the antenna coils, and consequently a static leak of considerable impedance is inserted.

The search coil itself is in turn also duly coupled to the amplifying system by means of a screened transformer. The screen of this transformer is grounded and eliminates the possibility of an out-of-balance electrostatic coupling, which is one of the causes of the insidious vertical component referred to above. This vertical component produces errors in the readings given and is due to the loops themselves tending to act as plain aerials instead of as directional frames.

Figure 7 shows the actual connection of the radiogoniometer and screened transformer and readily illustrates the manner in which the screen effectively insures that the amplifying system responds only to the currents flowing in the search coil circuit.

Amplifying Circuit

The amplifying device of this Marconi Bellini-Tosi direction finder consists of five stages of radio frequency amplification, a detector, and one-stage of audio-frequency amplification. The tubes employed in the radio and audio amplifications are the well-known Marconi V-24 type, whilst the detector is of the QX variety. The working current of the tubes is about 0.7 ampere each.

There is nothing of particular interest in this form of amplifier which embodies the usual radio frequency air core transformer wound on cylindrical ebonite formers.

"Sense" Finding

Plotting a polar curve for the current received by a rotating frame or the search coil of a Bellini-Tosi for every direction of the incident wave to such systems results in a "figure of eight," or bi-directional diagram. (Figure 8.)

It can be seen from this diagram that in the direction in which signals are at a maximum the receiving power of the antennas undergoes a very small change over a considerable angle, whereas the same change in angle in the direction in which the signals are a minimum gives a very large change in the signal strength, illustrating graphically the reason for the reading of minimum points when taking bearings.

This diagram, however, produces two minima and therefore is only useful for indicating the bearing of the incident wave from the system and does not denote its actual direction of origin.

To produce this actual direction a uni-directional or cardioid diagram is necessary and this can be obtained by combining the

effects of a vertical antenna which is equally sensitive to signals from all directions, with those of a loop.

Suppose a loop and a vertical antenna be both coupled to a common circuit in such a manner that the current induced in this circuit by the incident wave on the vertical antenna is exactly equal to that induced in the loop when the plane of that loop lies along the direction of propagation and is, therefore, at its optimum portion of reception. It will be seen that the electromotive force induced in the vertical antenna is either in phase or 180 degrees out of phase with the electromotive force induced in the frame according to the actual direction and consequently the two currents will either add or subtract giving one maximum and minimum for the whole 360 degrees. This addition and subtraction of currents for the whole rotation gives the cardioid diagram shown.

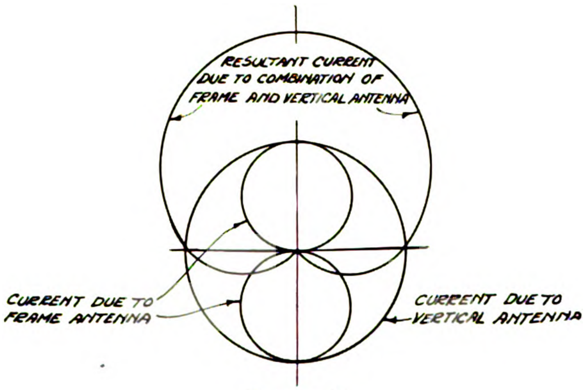

FIGURE 8

In the actual marine direction finder under discussion the tow loops are themselves made to function also as a vertical antenna as well as individual loops. This is accomplished by connecting the mid-points of the two loops thru a resistance to a tertiary winding of the afore-mentioned screened transformer. The resistance mentioned performs the double function of adjusting the phase and also the amplitude of the current. The value of this resistance is so chosen as to give the best minimum possible over the whole band of wave lengths covered by the transformer and its tuning condenser. Actually there is only one wave-length where perfect zero balance is obtained. Figure 9 shows the connections when the apparatus is functioning under this "sense-finding" arrangement.

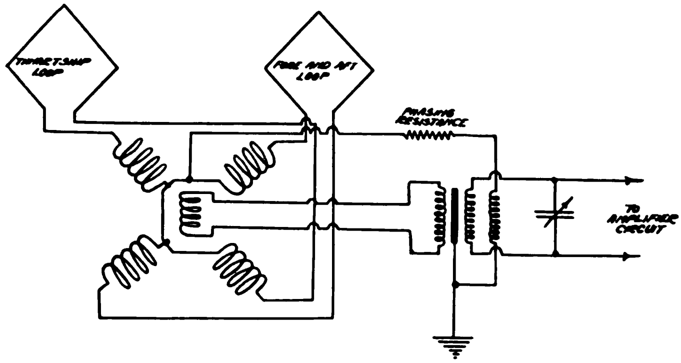

FIGURE 9

The sense minimum is never utilized alone in marine working as it is not usually sufficiently defined to guarantee accurate working and it is therefore used in conjunction with the more pronounced minima of the bi-directional diagram.

The actual apparatus (see Figure 10) is equipped with a switch which enables the operator to use the instrument in either the bi- or uni-directional functions which are known as the "DF" or "Sense" positions, respectively. There is a third position of this switch, known as the "Stand-bi" position, when the whole system functions purely as a simple receiving circuit and is used in searching for signals.

The complete circuit of this direction finder is shown in Figure 10.

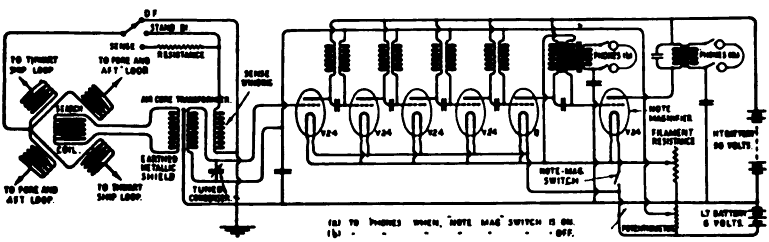

FIGURE 10—Two Small Variable Calibrating Chokers Are Omitted for Sake of Clearness. (These Are Actually Connected in Each Side of Fore and Aft Loop Circuit)

ERRORS

The errors that are likely to be encountered with the Marconi Bellini-Tosi marine direction finder can be divided into four classes.

Firstly, those errors associated with the antennas and the local external ships' rigging.

Secondly, inherent errors in the radiogoniometer instrument.

Thirdly, errors due to atmospherical conditions, which include the well-known sunrise sunset and night effect, and,

Lastly, errors due to coastal refraction.

The former two are governable by mechanical or electrical corrections whilst the latter two present a much more difficult problem, necessitating the taking of observations under special conditions.

Antenna Errors

Calibration Error

In equipping a radio direction finder antenna system on board ship, it is not possible to follow the process adopted in the case of the basic apparatus described, that is to say, two identical oriented receiving loops cannot be utilized.

The reason for this is due to the fact that the ship itself, being of metal construction with all its rigging, can be considered as an auxiliary receiving loop, lying in a vertical plane parallel to the keel of the vessel. This introduces an artificial loop parallel to the fore and aft loop, and therefore electrically coupled to it with the resulting mutual induction effect. This effect tends to improve the reception qualities of the tangible fore and aft loop. Naturally it is impossible to erect the loop outside the electrical coupling of this fictitious loop furnished by the ship's keel, and to overcome this difficulty the thwartship loop has to be made considerably larger than that of the fore and aft loop. This latter loop has in its electrical circuit two chokes, so that a final calibration adjustment can be made to render the reception qualities of the two loops equal.

Loop Tuning Error

The possibility of a loop tuning error is brought about by the fact that the two loops must necessarily be of slightly different sizes for reasons mentioned just above, so that the frequency of the two circuits regarded as two loops will be different. If one loop happens to be in tune with the wave being received and the other not, a very large error may result as the current induced in the loop that is in tune with that of the incident wave will be very much too strong.

This error, however, will not be present unless the frequency of the loops is quite close to that of the received wave. In order to overcome this difficulty, the wave lengths of the loops are kept well below 450 meters and it is for this reason that the

use of the special low capacity cable mentioned above is imperative.

Lack-of-Symmetry Error

The lack-of-symmetry error is caused by the electrical constants of the two halves of one loop measured from the apex to the mid-point being unequal. If this error is present it will result in the antenna coil, which should be giving say a zero current, producing a current which in turn will induce energy in the search coil and so introduce an error or tend to make the minima ill-defined.

The causes of this lack of symmetry are two-fold; the first of which is that due to unequal distribution of any electrical dimensions between the two sides of the loop, which result in unequal impedance in the two halves measured from the apex to the midpoint. This can be corrected by reconstructing the loops.

The second cause is that due to re-radiation from conducting portions of the structure and rigging of the ship which may have unequal effects upon the two halves of the loop. This is a difficult fault to discover and also to eliminate, and requires a careful planning of the antenna loops at the time of fitting.

In fitting a vessel it is essential that both the loops have a very pronounced apex so as to eliminate the alteration of the apex due to the pitching and rolling of the vessel which would otherwise result in the symmetry of the system being destroyed.

Instrument Errors

Vertical Component Error

One of the chief errors with regard to the instrument to be combatted is that known as the vertical component error. This error is brought about by superimposing the stray capacity coupling between the antenna coils and the search coils on the magnetic coupling. The effect of this stray capacity coupling is to distort both positions of zero result of coupling so that the two zero readings are not exactly 180 degrees apart, altho the line bisecting the angle between the observed zero is at right angles to the proper zero due to the magnetic coupling only. The presence of this error, however, is detrimental to rapid and accurate work and can easily be eliminated by the introduction of an earth shield between the windings of the transformer connecting the search coil with the tuning condenser, such as is described heretofore.

Quadrantal Error

This second instrument error is due to the fact that the coupling of the search coil and the antenna coil as a whole is not perfectly constant for all positions of the search coil relative to the antenna coils. By employing a search coil with a specially disposed winding, we can overcome the varying coupling effect so that errors due to this effect are no longer experienced in practice.

Errors Due to Atmospherical Conditions

(Sunrise, Sunset and Night Effect)

The taking of bearings by a radiogoniometer during sunrise and sunset is not desirable as it is during these periods that the well-known Heaviside layer effect is very prominent, with the result that electromagnetic waves are polarized to a certain extent and are tilted to such an angle as to produce a false directional effect in the whole system. This effect is also noted sometimes during the night, and whilst antennas of certain construction have been designed to overcome this difficulty it has not been possible at the moment to instal them on board ships, owing to the very poor receptive qualities of such antennas.

The usual manner of combating this difficulty is to take "large swing" readings during the night period, observing two points on the scale of equal sound intensity rather than to work to a fine minimum.

Night effect can be artificially produced by the transmitting of signals from an aeroplane carrying a trailing antenna, the electromagnetic waves so produced being of a polarized nature

Coast Line Refraction Error

It has been found by experience that radio bearings are distorted or thrown out of line if the line joining the transmitting station and the receiving direction-finding station cut a coast line at an angle of 15 degrees or less or if high land intervenes either close to the transmitting or receiving station.

The refraction of the electromagnetic waves may produce an error of as great as five degrees and is nor always constant in direction, altho for one particular locality this error for a given direction is found not to vary; this error will also alter considerably for very small changes of position.

In order to overcome this difficulty, navigators are warned to study the contour of the land surrounding the transmitting station, the bearings of which it is his desire to observe.

Special sketch charts are being slowly compiled of localities

where such errors are noticed, giving the arcs of good bearings observed of the known coast stations; a typical example of such a chart is shown in Figure 11.

Conclusion

In conclusion, it perhaps will not be out of place to detail one or two instances showing the use of the marine direction finder.

The S. S. "Stavangerfjord," proceeding on her first voyage from Christiania, after being fitted with a Marconi direction finder, received distress signals from the S. S. "Otta," which was drifting helplessly with her rudder-stock broken. From the position sent out by the "Otta," and accepted without question, it appeared she was 275 miles away, and the "Stavangerfjord" steered toward the position given. On arriving there, however, no trace of the "Otta" could be found, and observations with the direction finder indicated that she was sixty miles away. The course was altered accordingly, and the information given by the direction finder was found to be correct. Meanwhile the salvage steamer "Jason" had been sent out from Bergen to assist the "Otta," and the "Stavengerfjord" stood by. It soon appeared

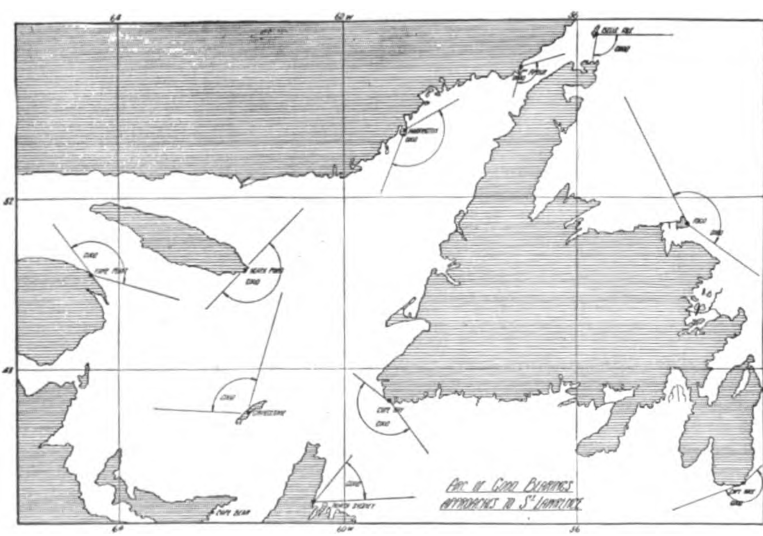

FIGURE 11

that the "Jason" was steering a wrong course, owing to her inability to take observations by the sun, and the "Stavangerfjord" piloted her to the right place by repeated directional observations.

During somewhat heavy storms in the North Atlantic the Norwegian steamer "Mod" was so badly damaged that she became practically a wreck, and for thirty-six hours the crew were huddled on deck without food. The captain sent out an "SOS" message, giving what he believed to be his position, but which proved erroneous. At least six vessels diverted their courses in an endeavor to render help, but no trace of the "Mod" could be found. For some time the British vessel "Melmore Head" was too far away to be of any assistance, but the captain kept in touch with what was happening, and when he found the "Mod's" signals getting stronger, he directed the radio operator to ascertain her position by means of his direction-finding apparatus. According to the reading thus obtained, the "Mod" was seventy-eight miles away from the position she herself had sent out, and in an entirely different direction. The captain of the "Melmore Head" placed his reliance on the direction finder and found it to be correct, arriving at the foundering vessel just in time to save twenty-three members of the crew before the "Mod" sank.

The rescue of the crew of the Norwegian steamer "Ontaneda" constituted another triumph for this latest development of radio science. A heavy gale had left the "Ontaneda" drifting helplessly with broken-down engines and listing at an angle of 50 degrees in a heavy sea. Her captain sent out the "SOS" signal for help, but in the thick weather he could get no observations of sun or stars, and had to estimate his position by dead reckoning. His calculations proved to be ninety miles out. Several vessels went to his assistance and steamed about near the position given without finding a trace of the "Ontaneda," but the S. S. "Fanad Head," by means of her direction-finding apparatus, discovered the true position of the vessel. She was nearer to the "Fanad Head" than to the ships which had originally steamed to her assistance, and the captain of the "Fanad Head" proceeded to the spot where he calculated the "Ontaneda" to be. Thus the radio direction proved to be the correct one, and the distressed sailors were rescued just in time.

An instance of the utility of the direction finding apparatus installed on board the White Star Liner "Megantic," is forthcoming in the case of the great assistance afforded to that vessel in locating the position of the U. S. S. "Charlot," after the latter struck an iceberg in dense fog and requested the "Megantic's" help. The commander relied implicity on the bearings taken, and the "Megantic" was only some ten miles distant when the "Charlot" advised that help was no longer required.

The master of the S. S. "Rosalind" reports that on his way to help the distressed steamer "Thyra," which was drifting rapidly before a strong breeze, he ran down to the position given by her, but could not find her anywhere. He then tested the direction-finding apparatus, called up the "Thyra" and got a bearing, ran straight on it, and proved the utility of this piece of apparatus. On the day following this incident, in a blizzard, the captain of the "Rosalind" directed the S. S. "Eastern Course" to his ship by means of the direction finder.

The salvage of the Norwegian ore-carrying steamer "Capto" was probably due entirely to the fact that its true position was plotted by the Marconi direction finder on board the S. S. "Montclare," when other vessels failed to locate its position from the directions sent out by the "Capto" itself. The "Capto's" wireless distress call, reporting the loss of a rudder and asking for assistance, was picked up by "Sachem" when the vessels were about 100 miles apart; and the "Sachem" immediately steered in the direction given. A heavy gale was raging, and altho the "Sachem" searched thoroughly, she could not locate the disabled ship until the Canadian Pacific liner "Montclare" determined the exact position of the two ships by means of its direction finder, and communicated the information to the "Sachem," which then came up with the "Capto." A tow of 750 miles to St. John's, Newfoundland, followed, ten days being taken to weather the conditions and reach harbor.

When on her first voyage with a Marconi radio direction finder on board, the Canadian Pacific Steamer "Metagama" ran into a dense fog off Belle Isle and was navigated by the aid of bearings obtained with the direction finder.

A case of another kind was that in which the transfer of a sick man from one ship to another was materially assisted by the use of a Marconi Wireless Direction Finder. On January 10th, when 1,700 miles west of the Franch coast, there was an accident on board the American steamship "Eastern King" and one of the seaman was injured internally. There was no doctor on board, and a message was sent out asking for a ship, with a doctor, bound for New York. This was answered by the Italian steamship "Conti Rosso," which was 300 miles away. Steaming directly towards each other, the two ships were due to meet at midnight, but owing to the heavy sea running, the captains mutually decided that it was unsafe to transfer the injured man at that time, and it was arranged that the ships should meet at 7 A. M. In the morning there was a dense fog, but the "Conti Rosso" was equipped with a Marconi direction finder, and by

its use the ships were able to find each other without delay. They met at 6.45 A. M., and the sick man was transferred to the doctor's care and eventually landed at New York.

SUMMARY: A description of the circuits, construction, installation, and use of the Marconi Bellini-Tosi marine radio direction finder is given. The method of determining "sense" as well as "line of direction" of a distant station is explained.

The various forms of errors in reading and the methods of reducing or eliminating these are discussed.

The paper concludes with a description of a number of cases where the Marconi direction finder has contributed to the safety of life at sea in stormy weather.

RECENT DEVELOPMENTS IN VACUUM TUBE TRANSMITTERS

By

B. R. Cummings

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The present paper may be considered as a continuation of the one presented before the Institute on April 4, 1923, by Mr. W. R. G. Baker.

Near the end of Mr. Baker's paper, reference was made to a 20-kilowatt transmitter which was then under development. I shall begin, therefore, with a description of this transmitter, a number of which have been built and installed, going into its circuit and design rather in detail, because it is felt that they indicate most clearly the present accomplishments in medium power vacuum tube transmitters.

Before discussing this, and other transmitters, however, I would call attention to several standard practices which were adopted with their development.

It will be recalled that the majority of tube transmitters built prior to 1923 were essentially "antenna oscillators," that is, no intermediate circuit was used. The wave length was established by the antenna circuit, and change in wave length by a change in the antenna loading with suitable changes in plate and grid coupling.

Such transmitters have been found to be unsuited to some classes of service, due to one or the other of the following reasons:

(a) As the efficiency of tube transmitter circuits increased, thru subsequent development, and the tube efficiency carried well above 50 percent, resulting in an approximation to a square wave plate current, it has been found necessary to utilize a circuit which would suppress the radiation of harmonics from the antenna.

(b) When used on antennas which are not rigidly secured in place, a swaying of the antenna is accompanied by a

*Presented before The Institute of Radio Engineers, New York, April 2, 1924. Received by the Editor, April 3, 1924.

change in the transmitted wave length. This effect is aggravated appreciably at the receiving station by the low decrement of the continuous wave signals.

A satisfactory solution to the suppression of harmonic radiation from the antenna has been found in the "Intermediate Circuit" transmitter which, in its elements, is shown in Figure 1.

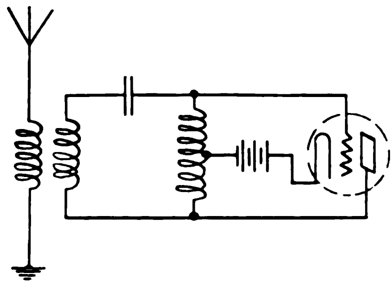


FIGURE 1

The intermediate, or "tank" circuit, as it is commonly called, is designed to have a circulating KVA. (kilowatt-amperage) many times that of the KW. (kilowatts) in the antenna circuit, usually in the order of from 12 to 15 times. It has been found that the suppression of harmonics is proportional to some degree on the ratio of intermediate circuit KVA. to antenna KW.

This circuit, when applied, brings us at once to the difficulties introduced by the so-called "drag-loop" effect encountered in intermediate circuit tube transmitters. Considerably more has been published abroad than in this country on this phenomenon. I will only include here a statement of the conditions under which the "drag-loop" is encountered.

If, for test purposes, the intermediate circuit transmitter is set up as shown in Figure 2, and an attempt is made to bring the secondary circuit into resonance with the primary by increasing the capacitance of C_2, it will be found that the frequency of the transmitter will slightly decrease as resonance is approached, and that in the neighborhood of resonance the frequency will suddenly change to an appreciably higher value, the extent depending, among other things, upon the degree of coupling between the circuits. Further increase in the capacitance of C_2 will cause a gradual decrease in frequency, which becomes constant at a value slightly higher than the original frequency.

This effect is illustrated clearly in Figure 3.

If C_2 is now gradually decreased, the sudden change in fre-

quency will again be noted, except that this time the change will occur at a smaller value of C_2. The resulting loop is the "drag-loop."

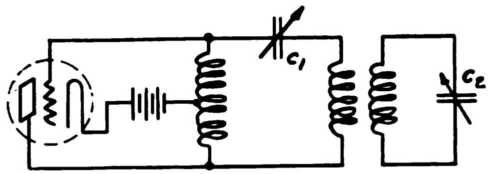


FIGURE 2

It is apparent that at values of C_2 lying between A and B, the circuit will oscillate at either of two frequencies and that operation is unstable. It has been found, in attempting to key a circuit of this kind, that the frequency set up would be either the higher or the lower one.

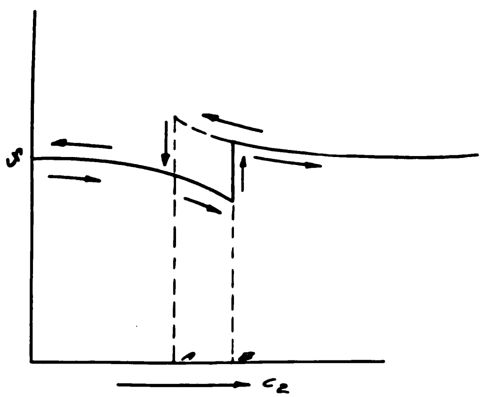


FIGURE 3

The cause of this phenomenon lies in the familiar laws of coupled circuits. It is encountered in spark transmitters when coupling is too tight, or when the gap does not quench properly, and all are familiar with the double hump in the frequency curve of such a transmitter. The lower frequency is established when the current in the intermediate and secondary circuits are in phase, the higher, when they are out of phase. These frequencies are lower and higher, respectively, than the natural period of the intermediate circuit.

In the vacuum tube transmitter there is no parallel to the quenching in a spark transmitter, so that the elimination of the drag-loop must depend on other considerations.

It has been found that the drag-loop may be eliminated in any of the three following ways:

(a) By having a very high circulating KVA. in the intermediate circuit. This solution is not economically practical.

(b) By secondary grid coupling, that is, coupling the grid to the secondary instead of the intermediate circuit.

(c) The use of a separately excited oscillator.

The third method has been adopted in transmitters in which an intermediate circuit is used. This has been done, primarily, because the separately excited oscillator also is of service in overcoming objection (b) to antenna oscillators, that is, the change in frequency with changes in antenna constants.

The separately excited oscillator, or the "master-oscillator" circuit, as it is more commonly called, has been adopted to hold constant output frequency and to facilitate the keying of the intermediate circuit transmitter, the intermediate circuit being provided, primarily, to suppress harmonic radiation from the antenna.

20-Kilowatt Transmitters

Outstanding among the transmitters which have been developed and built during the past year, are the 20-KW. vacuum tube transmitters built for the United Fruit Company for installation at a number of Central American locations for carrying the traffic of the United Fruit Company and the Tropical Radio Company.

Figure 4 shows the location of a number of these stations. It will be seen that they form a comprehensive communication system. The first three of these equipments have been installed and are in operation at Tegucigalpa, Honduras, at New Orleans, Lousiana and at Puerto Barrios, Guatemala. The installation of the remaining sets is now under way. Installations will also be made at Porto Barrios, Managua, and at Almirante.

The specifications laid out for the performance of these transmitters made it necessary to include positive means for the suppression of harmonics, and for maintaining constant frequency of output.

A schematic diagram of the rectifier of this transmitter is shown in Figure 5. It has an output of 2 amperes at 15,000 volts d. c. A 110- or 220-volt, 3-phase power supply supplies the primary of the plate transformer. The secondary is connected

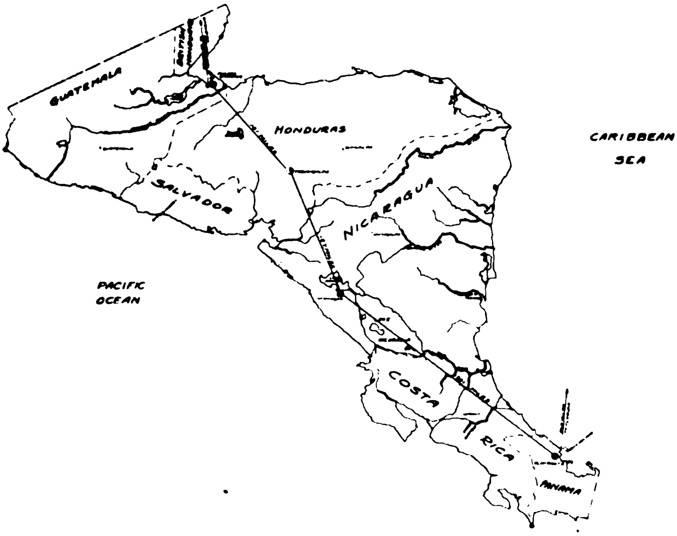


FIGURE 4

double Y with an interphase reactor between the common points of the Y's. A model UV-219 kenotron is connected in each leg of each secondary winding. The filaments are connected in parallel, and the load is taken from a mid point on the filament transformer, and a mid point on the interphase reactor. Smoothing condensers are placed across the output to decrease the ripple. These condensers are shunted by resistors, which are used to equalize the potential distribution across the condensers.

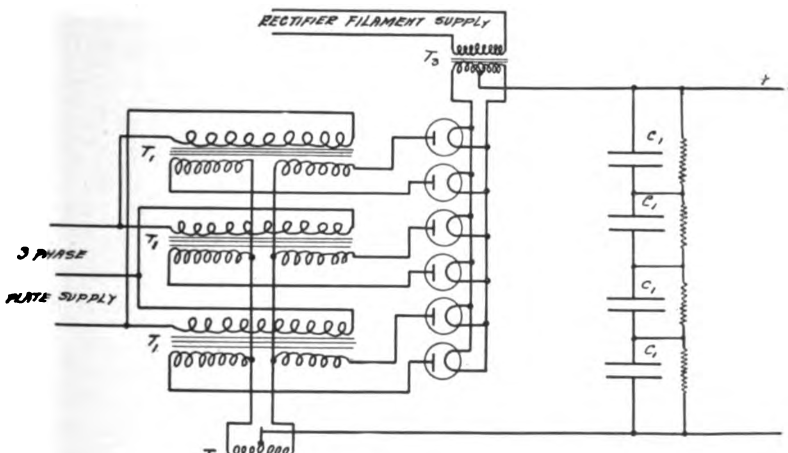


FIGURE 5

A schematic diagram of the complete transmitter is shown in Figure 6. The output of the rectifier is used to supply both the master oscillator and power amplifier, the plates of each of these tubes being operated at 15,000 volts, and the supply to them being thru suitable radio frequency choke coils L_1 and L_2.

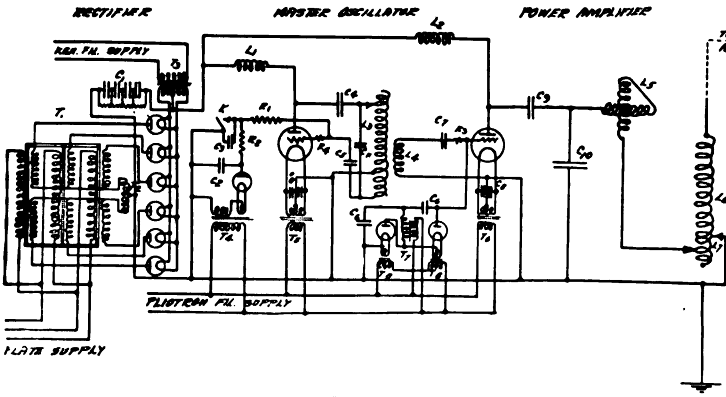


FIGURE 6

The master oscillator operates in a standard Hartley circuit, the output of which is fed to an intermediate circuit, the condenser of which is shown at C_{11}.

Grid biasing is accomplished by means of a kenotron rectifier, the circuit of which will be discussed in more detail later.

The grid of the power amplifier tube is coupled inductively to the coil system of the master oscillator, and is also biased by two kenotron rectifiers.

This method of biasing is of interest, and is shown more clearly in Figure 7.

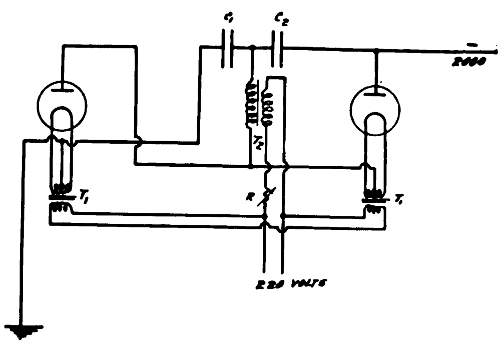


FIGURE 7

In this circuit, condensers C_1 and C_2 are alternately charged thru the kenotron in series with them, and the two condensers are in series with respect to the grid. The functioning of the circuit may best be understood by considering the rectifiers as functioning to keep the condensers C_1 and C_2 charged.

Referring again to Figure 6, the power amplifier operates into an intermediate circuit having a capacity C_{10}, and a variometer L_5 for tuning purposes. This circuit is coupled directly to the first down-lead of the multiple tuned antenna.

Keying is accomplished in the grid circuit of the master oscillator by opening and closing a circuit shunting the biasing rectifier. When the key is closed, the bias on the grid is controlled by grid leak R_1, but when the key is opened, a bias of approximately 2,000 volts is placed on the grid of the master oscillator.

All the equipment, except the outdoor tuning coils and their associated switching mechanisms, is installed indoors. The transmitters are built to cover a wave length range of from 2,500 to 4,500 meters. This range has been divided into two bands of from 2,500 to 3,300 and from 3,300 to 4,500. It will be noticed that in each case the wave length ratio is 1.3:1. Wave lengths within either of these bands are obtained by adjustments of the master oscillator tank variometer, the power amplifier, grid variocoupler, and the power amplifier tank variometer (4). The capacity values are not changed.

When changing from one wave length band to another, switches are thrown manually which change the capacitance of the master oscillator and power amplifier tank condensers, and change the coupling to the antenna tuning coil. During the tests of this equipment, it was found that the set gave stable operation over the required wave length range with a satisfactory safety factor in each direction.

The equipment, as built, is divided into the following mechanically independent units:

Control switchboard.
Kenotron rectifier.
Master oscillator panel.
Power amplifier tube unit.
Power amplifier tank circuit variometer.
Power amplifier tank circuit condensers.
Outdoor wave change and coupling change switches.
Outdoor tuning coils.

The power and control equipment consists of the following,

assuming that the main power supply is not generated at the station.

Starter and rectifier panel.
Filament motor generator set.
Operator's control switches.
Bias voltage control panel.

The filament motor generator is supplied primarily as part of the filament voltage regulator equipment, and provides means whereby a voltage regulator can effectively maintain constant voltage.

The voltage regulator has been included on a panel with the filament generator starting contactors, as illustrated in Figure 8. The voltage regulator shown in the glass cabinet to the right is of the vibrating type and maintains constant voltage by means of a vibrating contact which is actuated by a coil placed across the a. c. output of the generator. The vibrating contacts

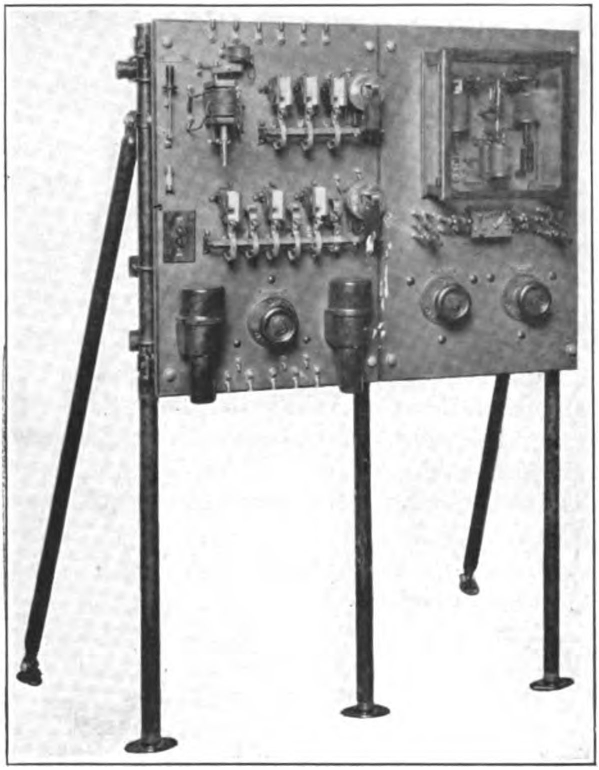


FIGURE 8—Starting and Regulating Panel for Use with Filament Lighting Motor Generator Set (A. C. Motor Driven) for 20 KW. Transmitting Equipments

act intermittently on the excitation of the filament generator. This regulator will maintain constant filament voltage over variations in supply voltage of from 95 to 115 volts, or its equivalent.

The control board is shown in Figure 9 and consists of the necessary instruments, controls, rheostats, and switches for the control of the set. Instruments are provided for reading total plate current, total power taken from the line, and the voltage impressed upon the kenotron, master oscillator, and power amplifier tube filaments. Two plate overload relays are mounted below the instruments, which, in the event of plate overload, open the holding coil circuit of the main line contactor, removing power from the set.

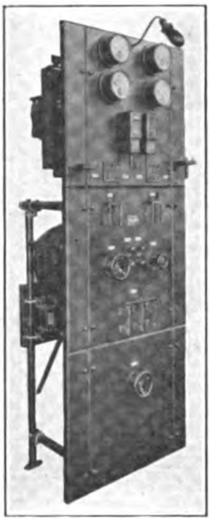


FIGURE 9—Service Panel for Use with 20-KW. Transmitting Equipment. Front View

The switches control the filament supply to the kenotron and power amplifier tubes, and rheostats are provided for filament voltage control and for plate voltage control. The latter rheostat is in series with the field of the power generator in those stations which generate their own power. For other stations, power change is accomplished by means of a switch, in the primary of the plate transformer which has three positions for each phase This switch is located in the same housing as the plate transformer and is oil immersed.

Figures 10 and 11 show more detailed views of the kenotron rectifier and its associated equipment. This rectifier is three-phase, double Y connected, and utilizes six UV-219 kenotrons. At full load the rectifier has an output of 2 amperes at 15,000 volts, with a ripple of 0.8 percent. At full load the rectifier efficiency is 92.7 percent, while the over-all efficiency, including filament consumption, is 83.6 percent. Other than the filaments, the losses in this unit are as follows:

Plate transformer....................	1,125 watts
Interphase transformer................	100 watts
Protective resistances.................	313 watts
Space charge..........................	790 watts

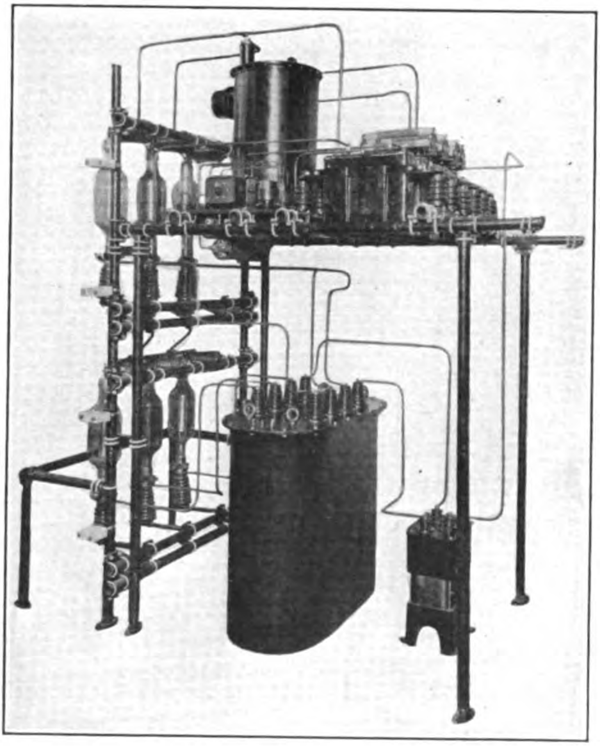


FIGURE 10—Kenotron Rectifier Unit for Use with 20-KW. Transmitting Equipment, Side View

This rectifier was given, among other tests, a heat run at full load for 12½ hours continuous operation, after which the maximum temperature rise in any of its parts was 48° C.

Referring to Figures 10 and 11, the plate transformer is shown

at the bottom of the rectifier, and in the rear of it the filament transformer for the kenotron tubes.

The large tank on the superstructure is the interphase transformer. In the rear of the interphase transformer on the superstructure are mounted the smoothing condensers and protective resistors. No controls are provided in this assembly, all of them being located on the service panel. On the superstructure (Figure 10), can be seen the small plate transformer, filament transformers, and model UV-217 rectifiers used for biasing the power amplifier grid.

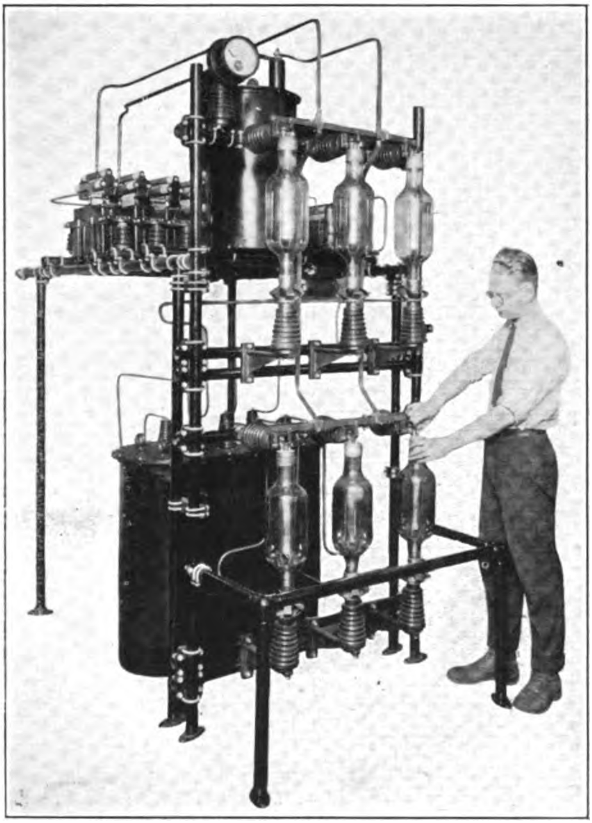


FIGURE 11—Kenotron Rectifier Unit (20-KW. Transmitting Equipment)

The resistances mounted across the smoothing contactors serve three purposes:

(a) As protective resistors to equalize the potential drop across the smoothing condensers.

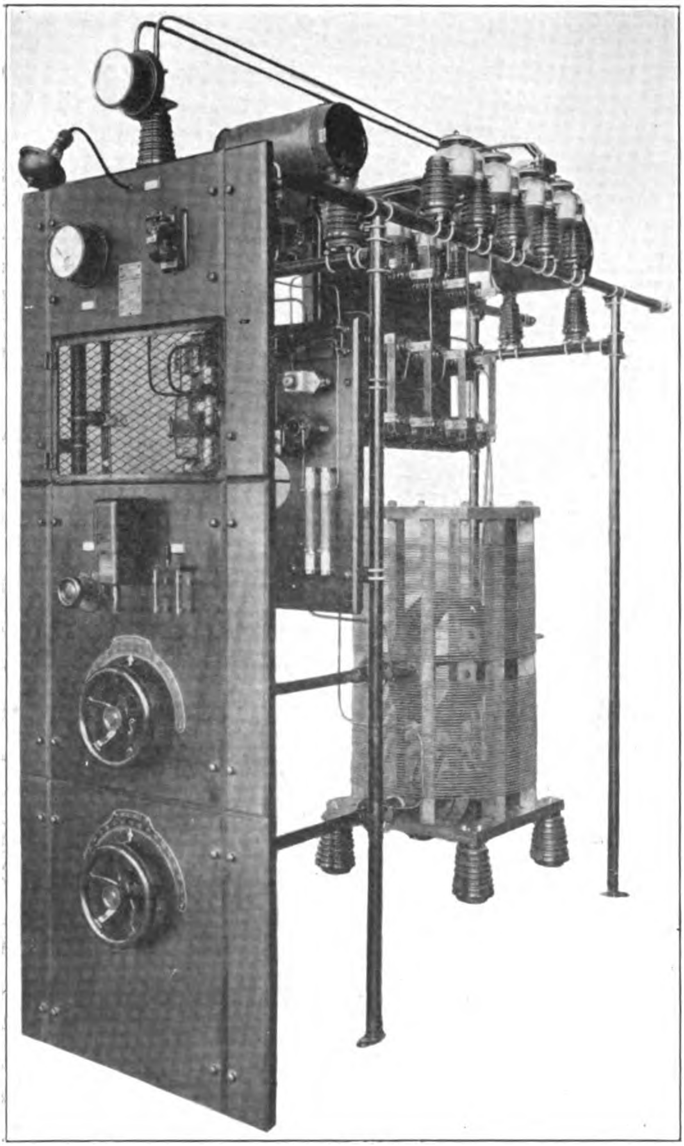


FIGURE 12—Master Oscillator for Use with 20-KW. Transmitting Equipment, Front View

(b) As a "dummy load" on the rectifier to improve its regulation.

(c) As a voltmeter multiplier.

The plate voltmeter is mounted on the insulator as shown on the top of Figure 11, and is in reality an ammeter con-

nected in series with the protective resistors, calibrated in volts.

Figures 12 and 13 indicate two views of the master oscillator assembly. The 1-kilowatt tube may be seen in Figure 13 mounted in the rear of the main panel protected by a mesh screen.

In the rear of the main panel is located the coil system of the master oscillator. The two rotors in this structure are provided for adjusting the wave length of the master oscillator, and adjusting the coupling between the master oscillator and the grid of the power amplifier.

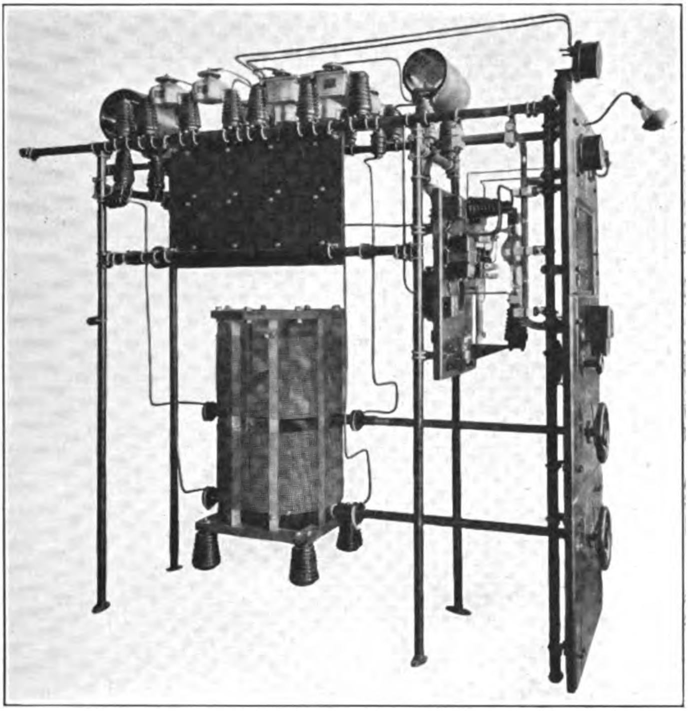


FIGURE 13—Master Oscillator for Use with 20-KW. Transmitting Equipment, Side View

The switches shown on the superstructure are provided for wave length change, and are used when throwing from one wave length band to the other.

In the upper right hand corner of the panel is mounted the keying relay, which is operated by a Morse key on the operator's table. The superstructure also contains the master oscillator tank condensers and plate choke coils.

The ammeter mounted above the panel on a porcelain insulator indicates the radio frequency current of the tank circuit of the master oscillator. The voltmeter on the panel indicates the voltage impressed on the master tube filament.

The contactors of the relay in the middle panel are connected in the ground lead to the filament transformer, in which position they protect the master oscillator tube from overload by opening the holding coil circuit of the main line contactor. The model UV-216 kenotron, which is used for providing bias for the master oscillator tube, can be seen on the sub-panel of Figure 13, together with its associated transformer.

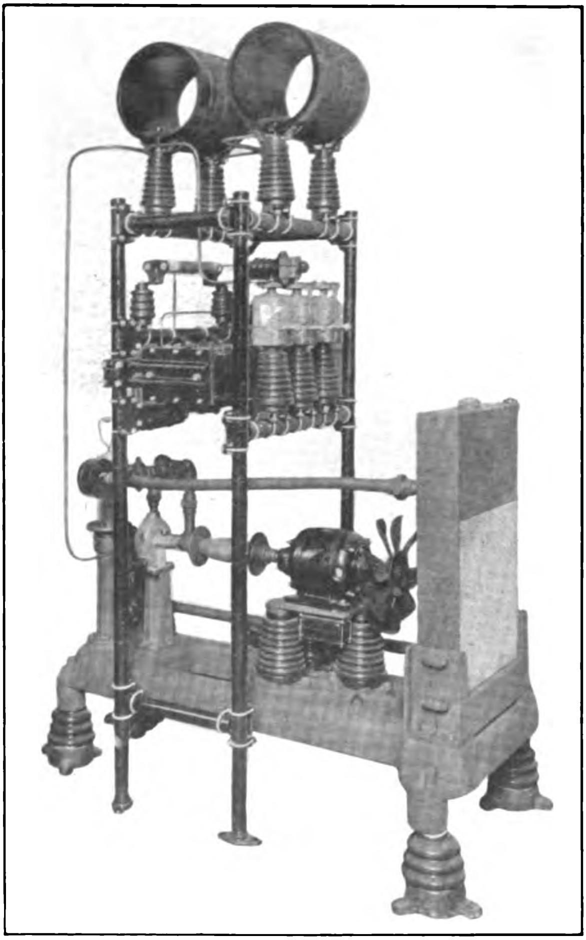


FIGURE 14—20-KW. (Radiotron UV-207) Water Cooled Tube Unit (Side View)

Figures 14 and 15 show front and rear views of the power amplifier tube and its associated cooling system and equipment. Figure 15 shows one of these tubes being lowered into the water jacket which forms part of the water cooling system.

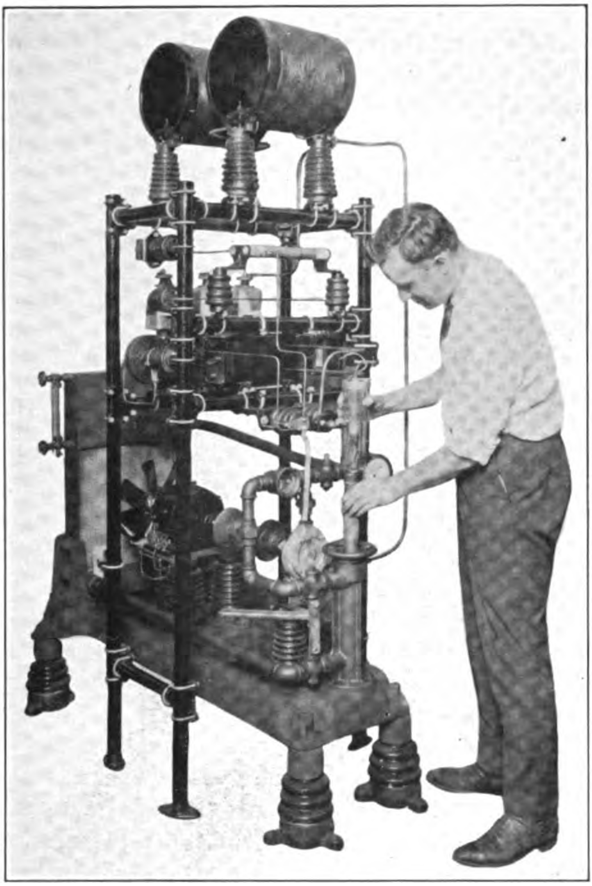


FIGURE 15—20-KW. (Radiotron UV-207) Water Cooled Tube Unit

This unit has been standardized upon for use with sets utilizing 20-kilowatt pliotrons, and has been supplied with various equipments other than those under discussion.

Since the anode of the pliotron is at a potential of 15,000 volts d. c. above ground in addition to the a. c. plate voltage, it is necessary to insulate the entire cooling system from ground for this voltage. For this reason the bed plate is substantially insulated from ground as shown in Figures 14 and 15. The cir-

culating system consists of a motor-driven pump, the water jacket of the tube, a radiator for cooling purposes, and necessary pipe work. Associated with the system is a flow indicator, a dial thermometer and a sight gauge on the radiator. The pump motor is also used to drive a cooling fan in the rear of the radiator. The motor and its associated circuit is insulated from the potential of the remainder of the cooling system by porcelain insulators shown in these figures, and by an insulating coupling between the motor and pump, which can be most clearly seen in Figure 14.

When removing a tube from its water jacket, the lever shown slightly above the coupling in Figure 15 is thrown which closes the inlet and outlet of the water system to the cooling jacket and also operates an auxiliary contact removing power from the set.

The cooling system as designed will dissipate 16 kilowatts, with a temperature rise of 15 degrees. It is imperative that the cooling water be not allowed to approach the boiling point, under which conditions the anode of the tube would not be completely in contact with the cooling water and hot spots would develop.

The system is also laid out so that it can be operated on the thermo-syphon principle, at reduced power. This provision was made to provide against any possible failure of the artificial cooling system. A by-pass is provided for the pump when so operated.

The superstructure shown in these figures contains the power amplifier, filament transformer, the plate-blocking condenser, grid leaks, and by-pass condensers.

High potential tests on the cooling unit show that it is capable of withstanding a potential of approximately 60,000 volts without flash-over.

The tank condensers for the power amplifier are shown in Figure 16. These condensers have capacitances, respectively, of 0.0021 and 0.0032 microfarads, and a power factor of approximately 0.03 percent. They will withstand a potential of 22,000 volts without flash-over. Their design is a flexible one and it is possible to build up condensers of this type with any number of plates, depending upon the requirements of the circuit.

Figure 17 shows one of the tuning coils built for outdoor installation. The concrete base on which these coils are mounted is built by the customer at the time of installation. These coils are wound with litzendraht conductor on porcelain forms.

Metal supports are provided at the base of the coil structure, which are broken by insulators to minimize losses. No terminals or taps are provided on these coils when built, the location of such taps being experimentally determined when the sets are installed,

at which time connections are made by suitably connecting taps for the various wave lengths.

Figure 18 shows the solenoid-operated high voltage disconnecting switch, six of which are used in each installation. Four are located at the first outdoor tuning coil, two for wave change and two for coupling change. Two are located at the second tuning coil for wave change. These switches are not in themselves suitable for outdoor mounting, but are mounted inside of protective housings erected at the tuning coils.

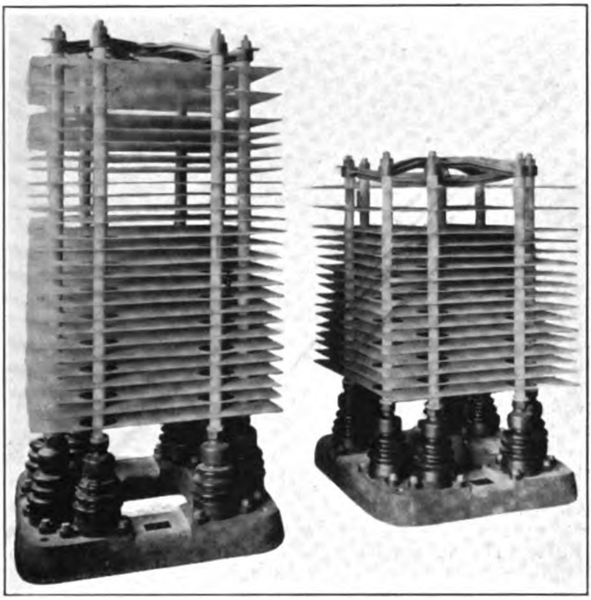


FIGURE 16—High Voltage Air Condensers

Figure 19 indicates the power amplifier tank circuit variometer. On this panel are mounted an ammeter indicating the current in the amplifier tank circuit, and a radio frequency ammeter indicating the current in the first down-lead. The latter instrument is operated from a thermo-couple located in the secondary of a radio frequency transformer, the primary of which is in the ground connection.

The switches in the lower right hand corner of this panel are provided for operating the outdoor switches of the tuning coils for wave length change. Two signal lamps are included, one above and one below these switches, to indicate the position of the outdoor contactors. The outdoor switches have asso-

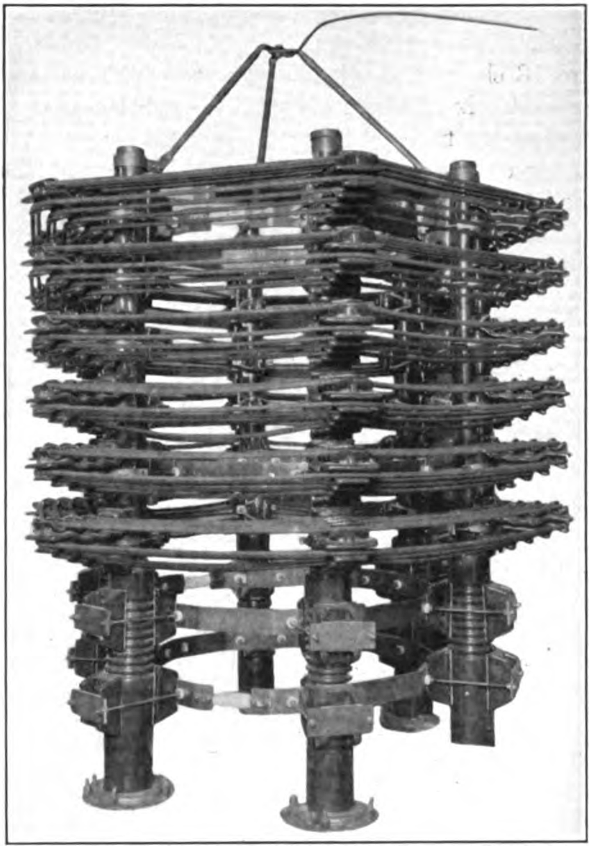


Figure 17—Tuning Coil for United Fruit Company, Radio Equipment

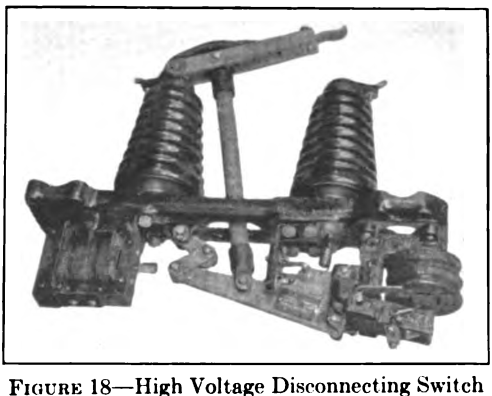


Figure 18—High Voltage Disconnecting Switch (Solenoid Operated), 25,000 Volts

ciated with them an auxiliary contact, which closes when the switch is in the closed position. Each of the signal lamps on this panel is in series with these auxiliary contacts on the outdoor switches so that positive indication is given in the station that all the outdoor switches have properly functioned when wave length is changed.

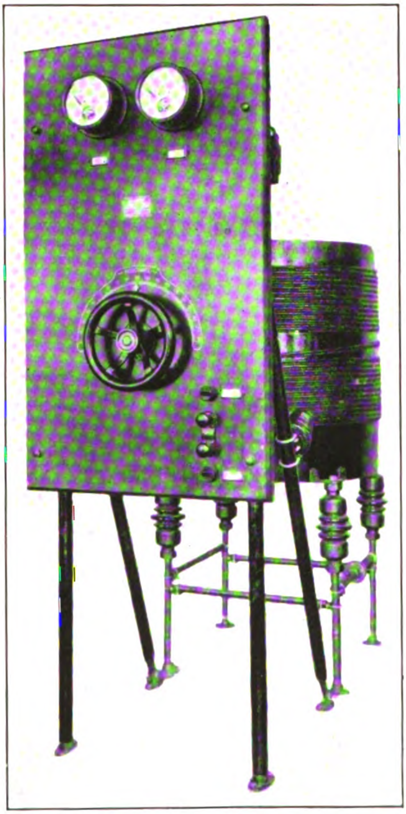


FIGURE 19—Intermediate Circuit Variometer, Front View

Figure 20 shows the operator's control panel, which is mounted on the operator's table and which includes all controls necessary for the normal operation of the set. The controls are operated in the sequence shown; power supply is connected by means of the plate contactor; the filament circuits are closed by the second contactor and the filament motor-generator set is started with the third contactor. The signal light at the top of this panel gives positive indication that the set is functioning.

Figure 21 illustrates the installation of one of these transmitters at New Orleans. This photograph illustrates the complete equipment, except for the filament motor-generator set and the outdoor coils and switches. The power amplifier tube unit is, unfortunately, not clearly visible in this picture.

Figure 21-A illustrates the installation of one of the outdoor tuning coils. The switches for wave change and coupling change are mounted in the housing shown to the right.

FIGURE 20—Operators' Control for Use with 20-KW. Transmitting Equipment

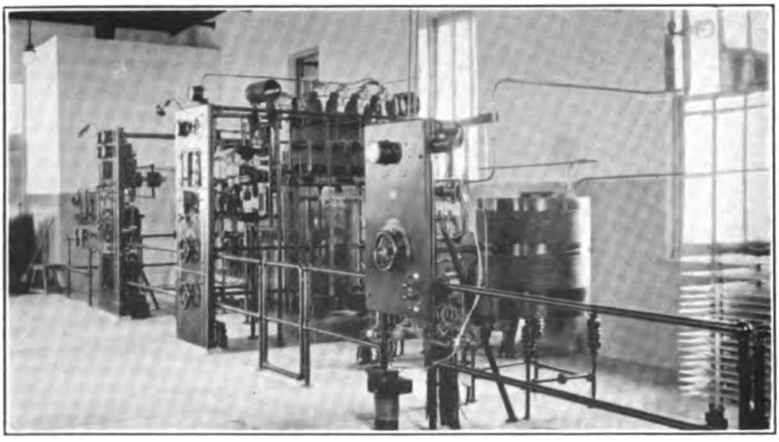


FIGURE 21

Complete reports are not available on the distances covered by thsee sets, altho the following results, obtained during preliminary tests, are of interest.

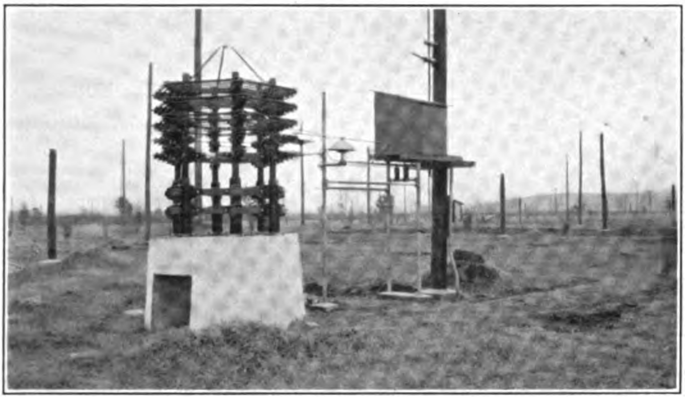


FIGURE 21-A

With the set at Tegucigalpa operating at medium power, and putting approximately 10 kilowatts into the antenna, average signals at New Orleans, a distance of 1,100 miles (1,780 km.), has an audibility of 400, daylight, and 700 night. At New York, a distance of 2,000 miles (3,200 km.), the average audibilities were 200 day and 600 night.

With 20-kilowatt input into the antenna at New Orleans, audibilities at Boston, Massachusetts, were reported in excess of 10,000.

10-KILOWATT SIGNAL CORPS TRANSMITTER

A 10-kilowatt telegraph transmitter has been built for the United States Signal Corps, which, while resembling the 20-kilowatt sets in many respects, has sufficient points of difference to warrant discussion. This transmitter is now installed in service at Fort Douglas, Utah. The circuit is substantially the same as that of the 20-kilowatt transmitters, altho the set differs fundamentally from the 20-kilowatt in the following points:

(a) The wave length range is 1,500 to 3,500 meters.

(b) Group wave change switches are provided.

(c) The characteristics of the power supply were such that no automatic filament voltage regulation was required.

(d) The set is designed to operate on a single tuned antenna.

The difference in the rating of these transmitters is mostly noticeable in the output of the rectifier. The rectifier for the 10-kilowatt transmitters, which is illustrated in Figure 22, gives an output of 21.6 kilowatts at 15,000 volts d. c., with a ripple of approximately 1 percent. It utilizes six model UV-218 kenotrons in a 3-phase, double Y rectifier, with an interphase transformer. the over-all efficiency of the rectifier, including filament consumption, is 81.2 percent. Power change is accomplished by taps on a switch in the plate transformer, giving plate voltage of 6,900, 10,700, and 15,000, respectively. The handle of this switch can be seen in the rear of the cover of the plate transformer, in Figure 22.

The master oscillator of this transmitter, shown in Figures 23 and 24, duplicates that of the 20-kilowatt transmitters, except that the coil system has been located externally, and its

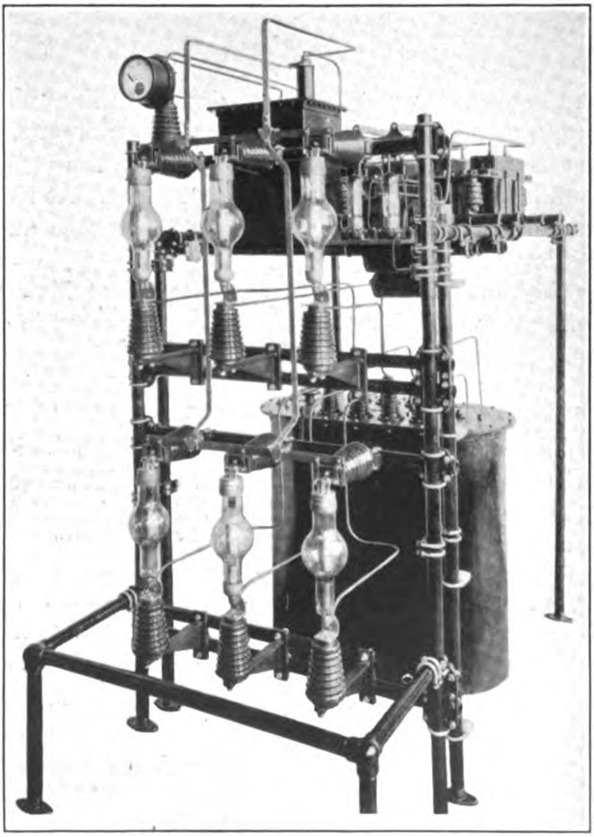


FIGURE 22—Kenotron Rectifier Unit (Signal Corps) Type RA-6

place taken by a three-bank wave change switch of novel construction.

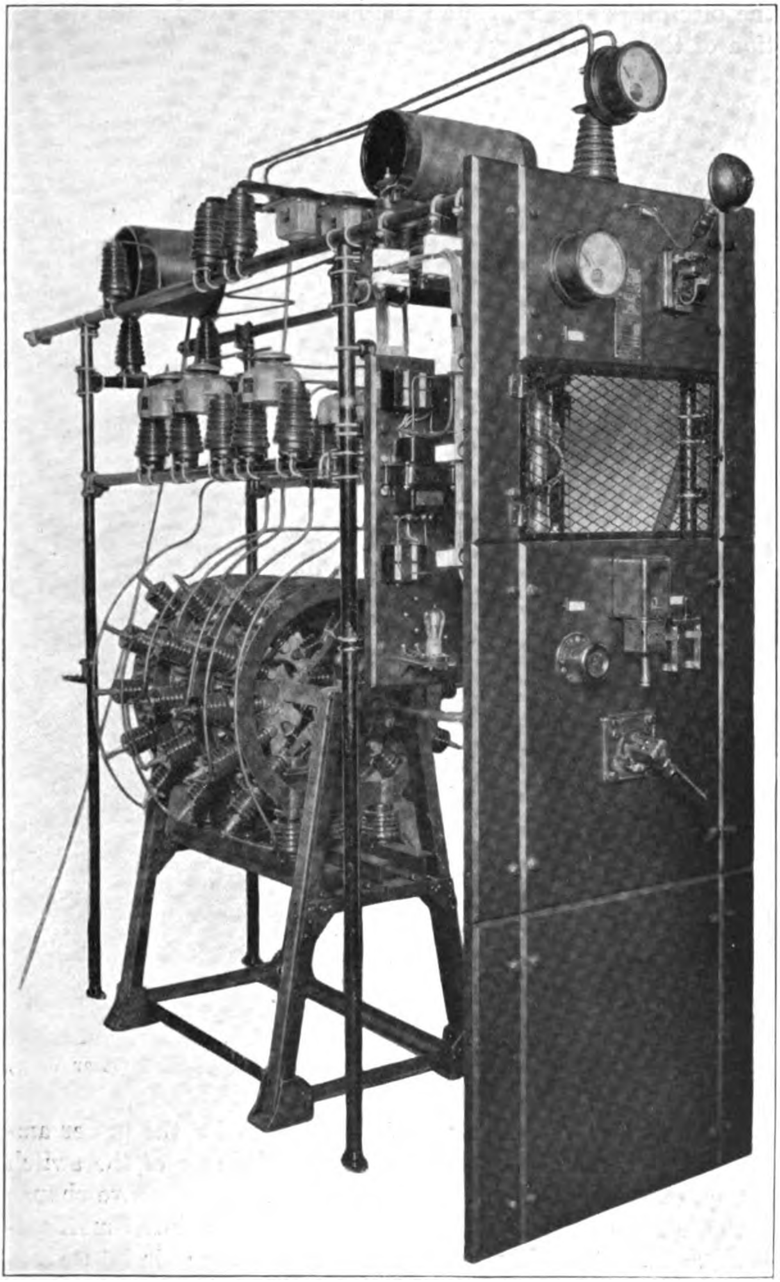


FIGURE 23—Master Oscillator Unit (Signal Corps) Type VO-2 (Front View)

Figure 25 shows the master oscillator coil system as an independent unit; the upper variometer varying the inductance of the intermediate circuit, and the lower one varying the excitation of the grid of the power amplifier.

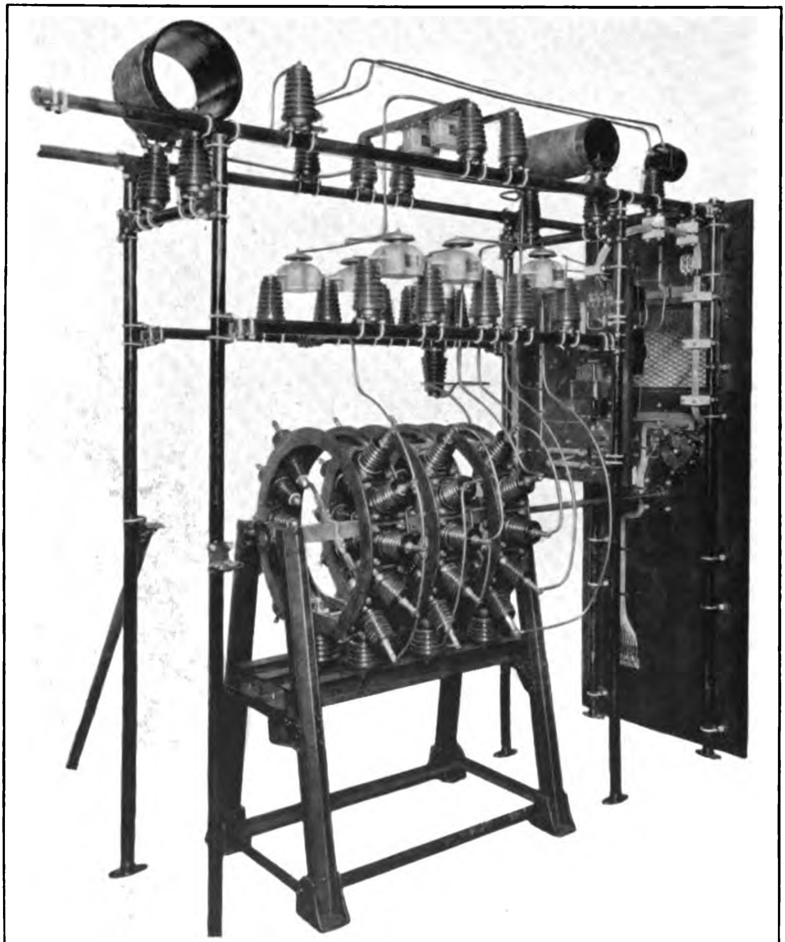


FIGURE 24—Master Oscillator Unit (Signal Corps) Type ID-9 (Rear View)

Figure 26 shows the wave change switch for the power amplifier and antenna, the largest bank in the rear of the switch being used to change the antenna loading. These wave change switches have been found to be extremely substantial mechanically, and are of a type which will probably be used in future sets requiring quick change in wave length.

Figure 27 shows the front of the same switch. The operating handle is so constructed that, when depressed, the switch is locked in position at the desired wave length. This panel also includes meters indicating antenna current and the intermediate circuit current of the power amplifier.

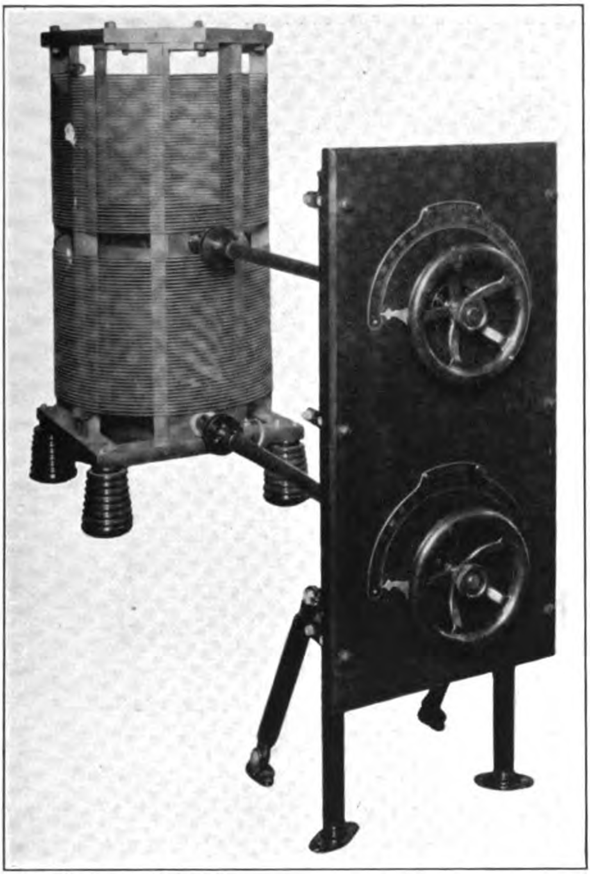


FIGURE 25—Master Oscillator Inductance (Signal Corps) Type ID-9

The loading coils for the antenna are shown in Figure 28. They are wound on treated wood supports to minimize losses.

Figure 29 shows the power amplifier intermediate circuit condenser, of the same type as those used in the 20-kilowatt transmitters, altho differing in the assembly. This condenser is rated at 22,000 volts. The three sections have capacitances, respect tively, of 0.0005, 0.0008, and 0.0012 microfarads. With these

three values of capacitance, combinations can be used for the five wave lengths as follows:

Wave change number 1............0.0012 microfarads
Wave change number 2............0.0013 microfarads
Wave change number 3............0.0017 microfarads
Wave change number 4............0.0020 microfarads
Wave change number 5............0.0025 microfarads

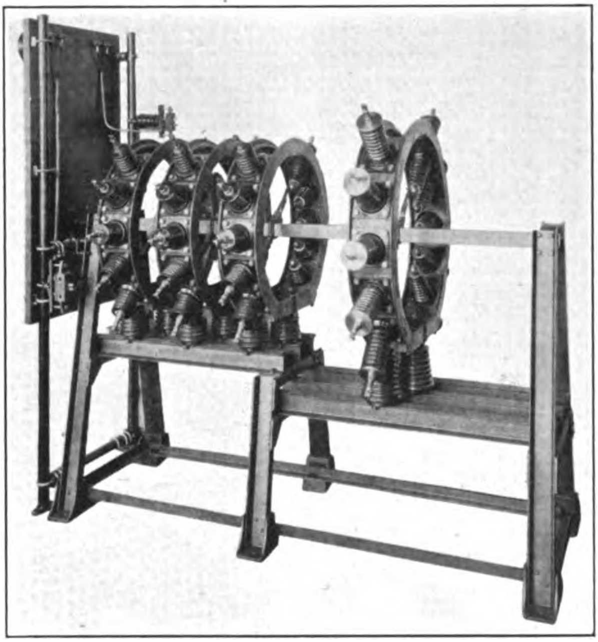


FIGURE 26—Wave Change Switch (Signal Corps) Type SW-83 (Side View)

The changes in connections to accomplish these capacitance steps are made by the power amplifier wave change switch.

The efficiency of this complete transmitter, based on the input from the line and the input in the antenna, including all filament consumption, is 56.5 percent at 1,700 meters.

Figure 30 is shown as a matter of interest to indicate the wiring on the service panel of one of the medium power transmitters. All wiring on these panels is of lead-covered conductor, to provide adequate shielding. The Radio Department has adopted the practice of photographing the first panel to be wired, and then using such photographs in the wiring of duplicate equipments. This procedure has been found much more

efficient than the previous practice of making actual drawings of the wiring.

Figure 31 is another photograph taken primarily for wiring purposes.

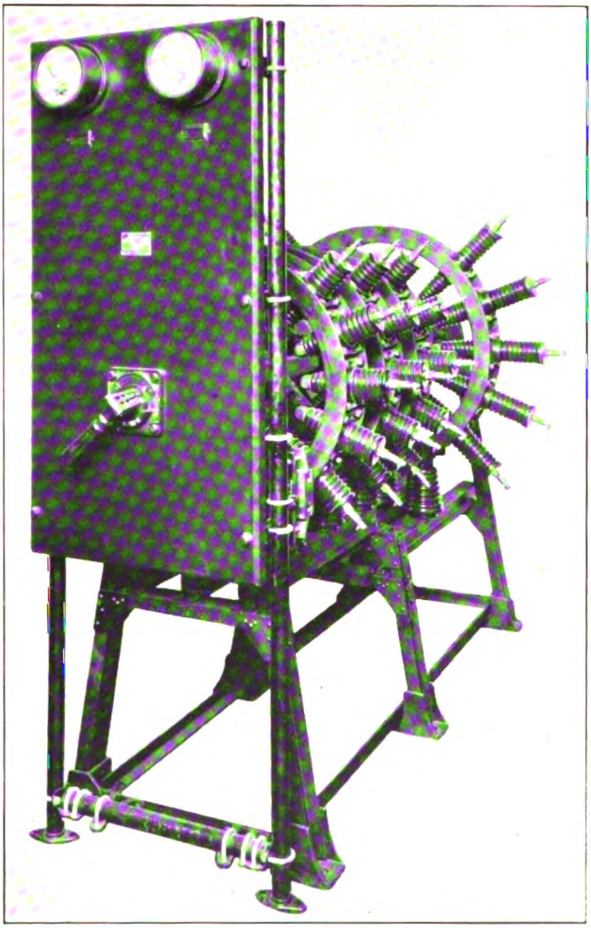


FIGURE 27—Wave Change Switch (Signal Corps) Type SW-83 (Front View)

In connection with the development of the transmitters first described, investigations were made to secure data upon which to base closer computation of the predicted performance of transformers and reactors, particularly filter and modulation reactors. An accurate, economical design of such reactors which carry an alternating current superimposed on a direct current, has been

made possible by careful investigation of "incremental" and "decremental" permeability. These investigations have brought to light much useful data for the design of this equipment which was not hitherto available, and it is hoped that the results of the tests will be made the subject of a separate paper.

FIGURE 28—Antenna Loading Coils (Signal Corps) ID-6 and ID-7

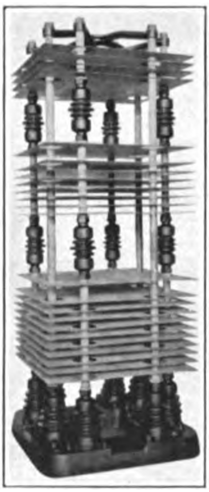


FIGURE 29—Intermediate Circuit Condensers (Signal Corps)

KEYING CIRCUIT

After a considerable amount of development work and service tests with various types of keying circuits, the one shown

in Figure 35 has been adopted for low power vacuum tube telegraph transmitters. This figure shows the essential elements of the keying circuit without reference to the oscillating circuits.

FIGURE 30 — Service Panel, 10-KW. Telegraph Set (Back View Showing Wiring)

FIGURE 31—Service Panel 20-KW. Telegraph Set (End View Showing Wiring)

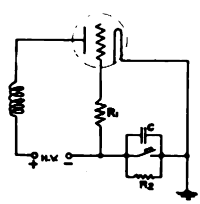


FIGURE 35

The filaments are grounded as usual. The grid, or grids of the tubes, are returned to the negative side of the high voltage supply with suitable grid leak resistors. The key is placed between the negative side of the high voltage supply and ground, and is shunted by a condenser C and a resistance R_2. The function of the condenser is merely to absorb sparking at the key contacts. The function of the resistance R_2 is to place a definite bias on the grids of the tubes when the key is up. In this position a small current, in the order of several milliamperes, flows thru resistance R_2, placing a bias of several hundred volts on

the grids of the tubes, resulting in a very positive keying system. This circuit is utilized in the majority of the medium power transmitters about to be described.

SCR-136 Mule Pack Transmitter

The manufacture of a vacuum tube mule pack transmitter was completed for the United States Signal Corps. The preliminary development of this transmitter was carried on by the Signal Corps at Camp Alfred Vail, New Jersey, and the further development and design was fundamentally the carrying out of the work which was started there.

The building of this set incidentally included the development of a gas engine driven generator, which is illustrated in Figure 36. This figure illustrates the power equipment in its carrying position, in which the generator is mounted apart from the gas engine.

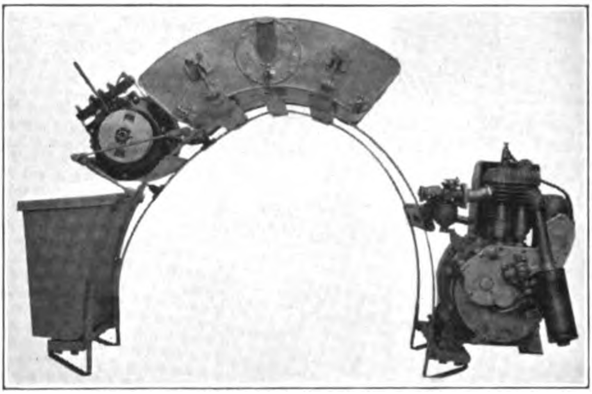


Figure 36—Mule Pack Power Unit for United States Signal Corps

The gas engine is a 4½ horse-power unit, and weighs 64 pounds (29 kg.). It is equipped with a governor, which provides 5 percent regulation from no load to full load. The engine runs at a speed of 2,500 revolutions per minute, and drives the generator directly at that speed.

The complete power equipment shown in this figure, including the gas engine, generator, gasoline tank, gasoline and tool box, weighs 250 pounds (118 kg.).

The generator, under full load, delivers 0.6 amperes at 900 volts, and 15 amperes at 11 volts. The power unit is rated at one hour's continuous operation.

Figure 37 shows the gas engine-generator unit assembled for operation. The coupling between the generator and the gas engine is accomplished by a single thumb-screw. The power equipment is of course removed from the mule's back for operation.

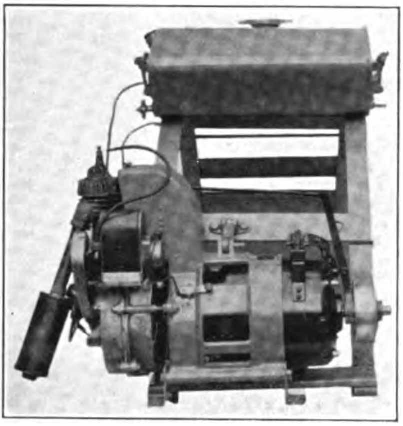


FIGURE 37—Mule Pack Power Unit for United States Signal Corps

The transmitter is completely housed in a weather-proof case for transportation, and provided with a vent in the top of the cabinet, which is closed in this condition.

Figure 39 shows the transmitter in operating condition. The vent in the cover automatically opens when the cover of the transmitter is lowered. The vent was found to be essential to assist the dissipation of heat generated by the tubes.

The weight of the transmitter unit alone is 43½ pounds (19.8 kg.). The containing case weighs 40 pounds (18.2 kg.), making a total weight of 83½ pounds (38 kg.). A circuit diagram of this transmitter is shown in Figure 40. It utilizes four 50-watt tubes; one as a master oscillator, one as a speech amplifier, one as a modulator, and one as a power amplifier.

The master oscillator tube is used in a Colpitts circuit, the wave length of which is controlled for the entire wave length range of the set by means of variometer L_7. The master oscillator is coupled to the grid of the power amplifier by means of the condenser C_4. The grids of the power amplifier and the modulator tubes are biased by means of a potentiometer, R_2 in the grid circuit of the power amplifier tube. Coupling between the speech amplifier and the modulator is accomplished by an iron core transformer T_2.

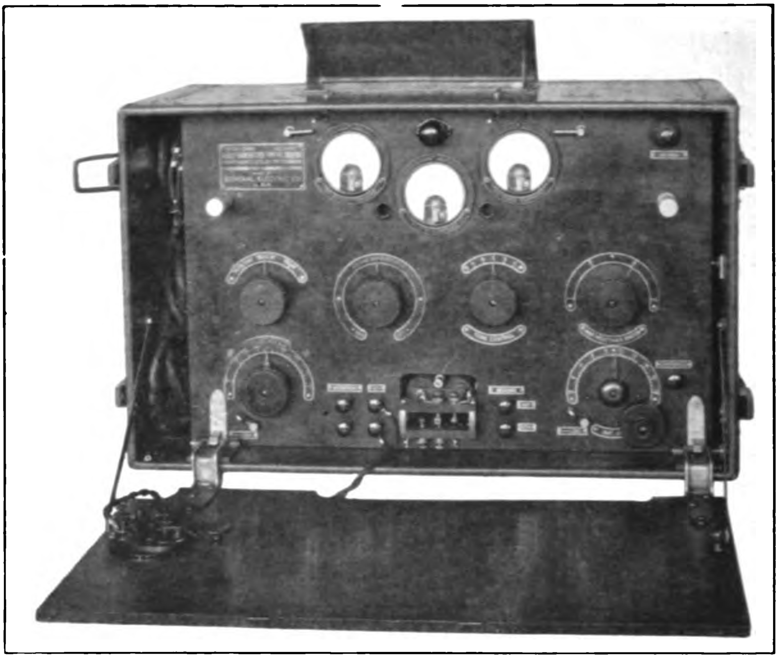


FIGURE 39—Mule Pack Set Radio Transmitter-Type SCR-136 (Front View Open)

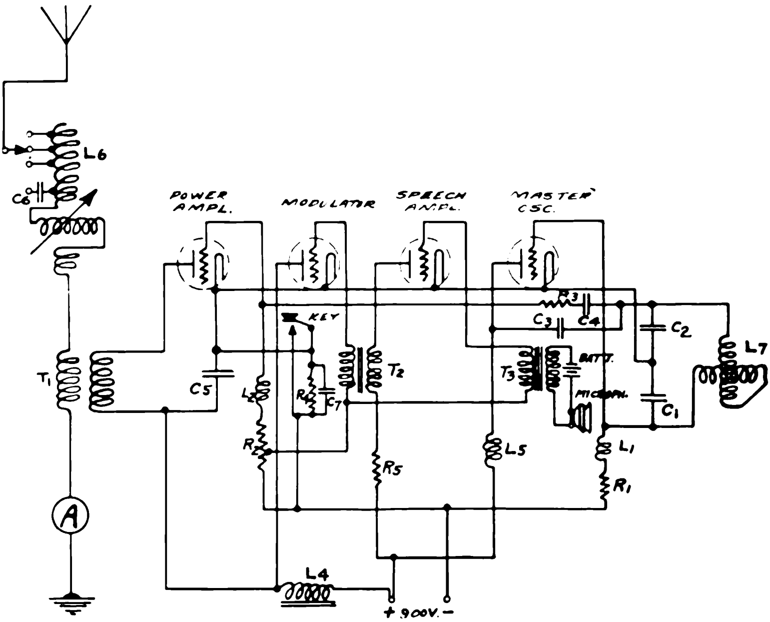


FIGURE 40—Schematic Diagram of Connections Radio Transmitter SCR-136

The transmitter is built for operation on telephone, continuous wave telegraph, and tone telegraph. Connections for the latter are not shown in the accompanying diagram, but they consist simply in the substitution of a small alternator for the microphone transformer. This alternator is of special construction, and is driven by a 6-volt battery. A rheostat in its field permits a selection of any one of five tones.

SCR-132 Transmitter

A second transmitter, known as the SCR-132, was built in co-operation with the Signal Corps Laboratory at Camp Alfred Vail, New Jersey. The circuit, as used, was established fundamentally by the Signal Corps Laboratory.

The circuit of this transmitter is shown in Figure 41. It utilizes a total of seven tubes; three model UV-204-A (VT-22), and four Model UV-211 (VT-4-B). One of the latter is used as a master oscillator, two as speech amplifiers and one as an intermediate amplifier. Two of the former are used as modulators and one as a power amplifier.

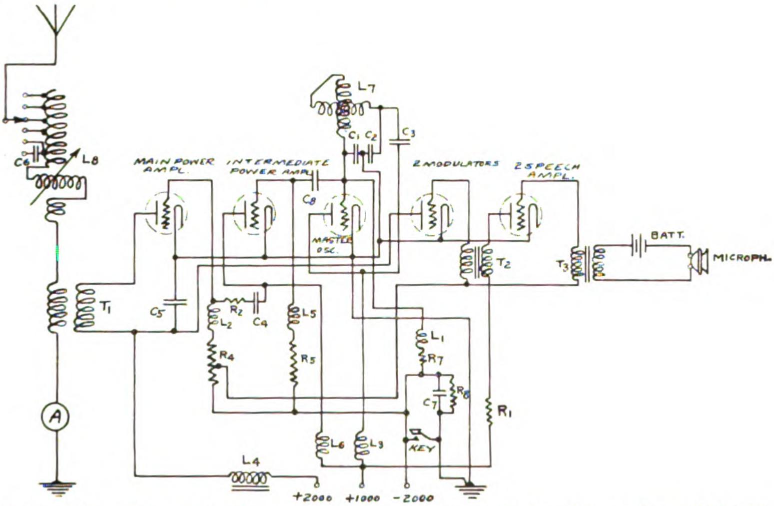


Figure 41—Schematic Diagrams of Connections Radio Transmitter BC-127

A Colpitts circuit is used for the master oscillator. The complate wave length range of the set, from 840 to 1,930 meters, is covered by the single variometer in this circuit.

In this, as well as in other medium power transmitters, the Colpitts circuit has been used for the master oscillator for two fundamental reasons: firstly, because of the mechanical sim-

plicity of constructing a variometer as compared to variable air condensers, and secondly, because of the fact that the circulating KVA. in the Colpitts circuit increases with increase in frequency, which is conducive to stable operation.

The power amplifier received its grid excitation thru a capacity coupling C_4 to the plates of the intermediate power amplifier. Power is delivered to the antenna circuit from the plate of the power amplifier by means of a closely coupled antenna transformer. Sufficient reactance exists in the plate winding of this transformer to limit the power amplifier d. c. plate current when the antenna circuit is detuned or open. Audio modulation is obtained by two speech amplifiers and two modulating tubes, operating in a modified Heising circuit.

The intermediate power amplifier is provided more as a means to insure constant frequency than to provide additional amplification. The output of the master oscillator would be sufficient to excite the grid of the power amplifier, altho the load would be such that the frequency would not be constant, particularly in view of the tight coupling in the antenna transformer.

The plates of the 50-watt tubes are operated at 1,000 volts thru suitable radio frequency chokes. The plates of the modulator and power amplifier tubes are operated at 2,000 volts.

To obtain tone telegraph, a small alternator is put in circuit in place of the microphone transformer.

Keying is accomplished by opening and closing the ground connection to the grids of the master oscillator, the intermediate amplifier, and the power amplifier tubes.

The equipment is portable and is shown in Figure 42 in its carrying conditions. The arms provided for carrying are utilized as legs when the equipment is put into operation as shown in

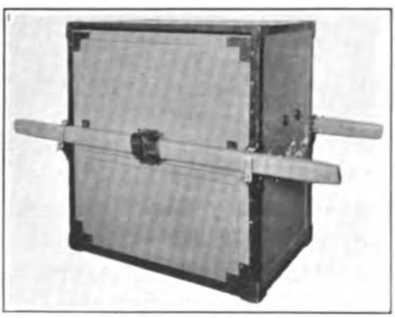


FIGURE 42—400-Watt Portable Radio Transmitter SCR-132 (Closed for Transportation)

Figure 43. The lower half of the cover forms a desk for the operator, and the upper half a partial shield against the elements.

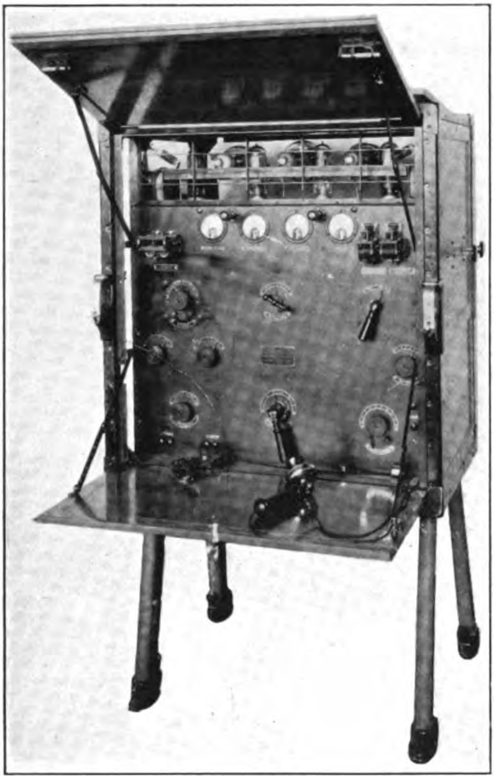


FIGURE 43—400-Watt Portable Radio Transmitter SCR-132 (Front View)

Figures 44 and 45 show end and rear views of the transmitter. Provision is made so that the transmitter may be remotely controlled for either telegraph or telephone communication. Transfer to remote control is made by means of a switch on the transmitter panel.

POST OFFICE AIRCRAFT EQUIPMENT

For installation on the planes of the Air Mail Service of the Post Office Department, there has been built a complete radio transmitting and receiving equipment. The design of this equipment includes many mechanical features, making the equipment extremely rugged and simple to operate.

The transmitter utilizes a total of six Model UV-203 50-watt

radiotrons; one as a master oscillator, one as a speech amplifier, two in parallel as modulators, and two in parallel as power amplifiers. The normal continuous-wave output of the transmitter is 150 watts.

The transmitter has a wave length range of from 190 to 290 meters.

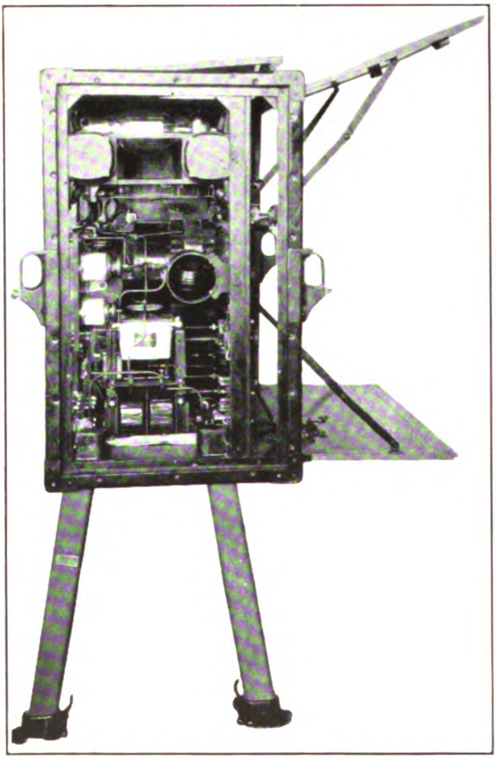


FIGURE 44—400-Watt Portable Radio Transmitter SCR-132 (Left Side Panel Removed)

Figure 46 shows schematically the circuit utilized in the transmitter. The master oscillator circuit is of the Colpitts type. With this arrangement, a calibrated variometer in the master oscillator circuit is set at the desired wave length. The antenna circuit is then tuned by the antenna variometer. This method of tuning is extremely simple and no skill is required to secure maximum output. Failure accurately to tune the transmitter removes load from the tubes.

The output of the microphone is used to actuate the speech

amplifier. These tubes operate as a resistance-coupled amplifier and are coupled to the grids of the modulator tubes.

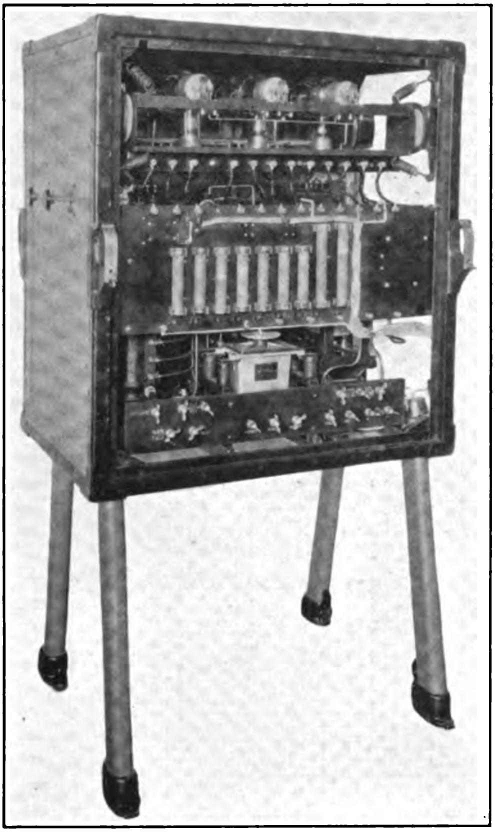


FIGURE 45—400-Watt Portable Radio Transmitter SCR-132 (Rear Panel Removed)

The power for the equipment is obtained from a dynamotor. This dynamotor has an output of 0.65 amperes at 1,000 volts, and is operated from a storage battery. The transmitter proper is divided into two units, that is, the transmitter proper and the control unit. The transmitter, which is shown in Figure 48, has only a single control, that is, the variometer in the master oscillator circuit. The assembly is supported on a duralumin frame, with a panel of insulating material. The tube sockets are rigidly mounted on the upper side of the transmitter. When installed, the unit is supported by eight straps and helical springs, the latter acting as a cushion to protect the tubes from vibration

and shock. Where space for installation is limited, the transmitter may be mounted back of the operator or out of sight. since the controls are mounted in a separate unit.

The control unit, illustrated in Figure 49, is designed with the intention of mounting it under the pilot's seat in the average installation. It contains a send-receive switch and the antenna variometer. The send-receive switch is operated by a substantial vertical handle and performs the following functions:

(a) Starts and stops the dynamotor.

(b) Transfers the antenna from transmitter to receiver.

(c) Opens and closes the filament circuits of the transmitter and receiver.

(d) Controls the keying circuit.

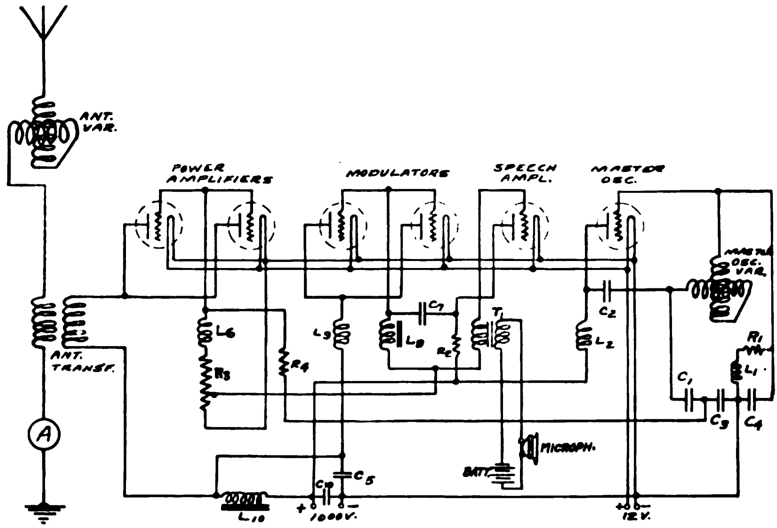


FIGURE 46—Schematic Diagram of Connections, Post Office Aircraft Receiver Unit

The antenna variometer, in series with the antenna, has sufficient range to tune the antenna circuit to any wave length covered by the master oscillator.

A telephone jack is included in the control unit, making a convenient connection for the operator's microphone.

A new antenna reel has been designed for this equipment, and is illustrated in Figure 50. The operator may release the handle of this reel at any time, and it will lock in place automatically. This prevents the loss of the antenna, which sometimes occurs with reels of other types that require a clamp and knob to hold them in place.

The receiver supplied with this equipment, which is illustrated in Figure 51, is of the super-heterodyne type, having the same wave length range as the transmitter. The inherent selectivity

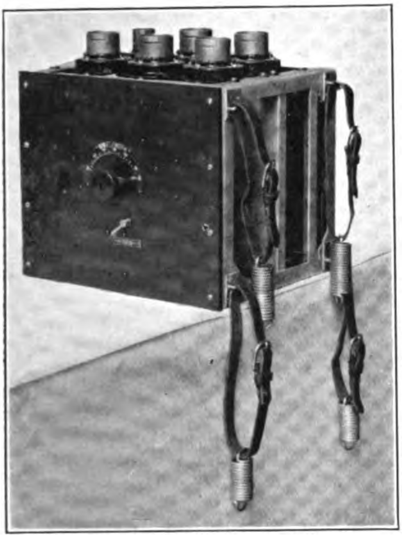


FIGURE 48—Radio Transmitter for Air Mail Plane (Front View)

FIGURE 49—Radio Control Box for Air Mail Plane

of this receiver, together with the practical elimination of engine noises by the use of several stages of radio frequency amplification, make the receiver well adapted to aircraft use. The receiver utilizes a total of seven UV-199 radiotrons as follows:

1 Radio frequency oscillator.
1 Radio frequency detector.
3 Stages of intermediate frequency amplification.
1 Stage of audio frequency amplification.

Its circuit is shown schematically in Figure 52.

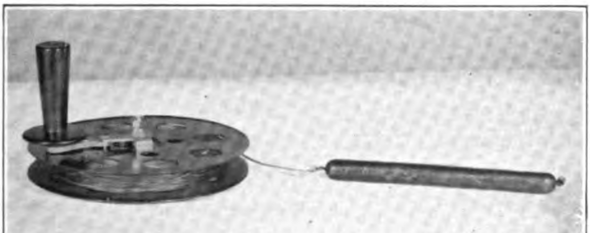


FIGURE 50—Antenna Reel and "Fish" for Air Mail Plane

Installation of this equipment for service tests on a plane of the Air Mail Service, is shown in Figure 53.

The transmitter is installed in the fusilage back of the operator. In this picture the top of the fusilage has been removed.

The receiver and antenna relay can be distinguished in the front of the operator, the former being installed under the control board of the plane. The control unit is mounted under the operator's seat, and is not easily seen in this photograph.

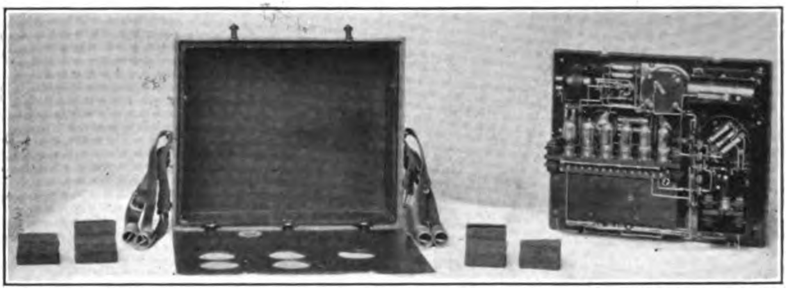


FIGURE 51—Radio Receiver—Air Mail Plane (Disassembled)

The service tests of this equipment have been very encouraging. The set has been operated by pilots with no previous experience in radio, and distances have been covered far in excess of the requirements of the set. The transmitter was required

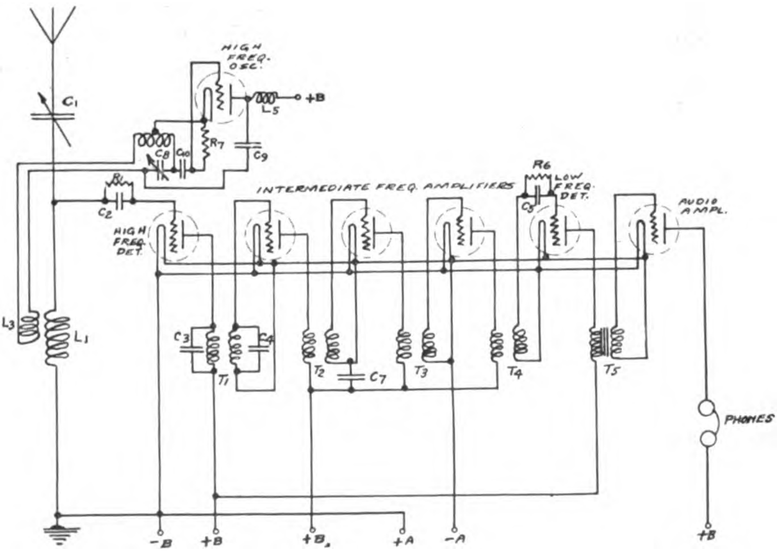


FIGURE 52—Schematic Diagram of Connections, Post Office Aircraft Receiver Unit

to have a range of 100 miles (160 km.) under specified conditions. This range was exceeded in the first service flight between Schenectady and New York, a distance of approximately 150 miles (240 km.), during which very satisfactory communication was maintained thruout.

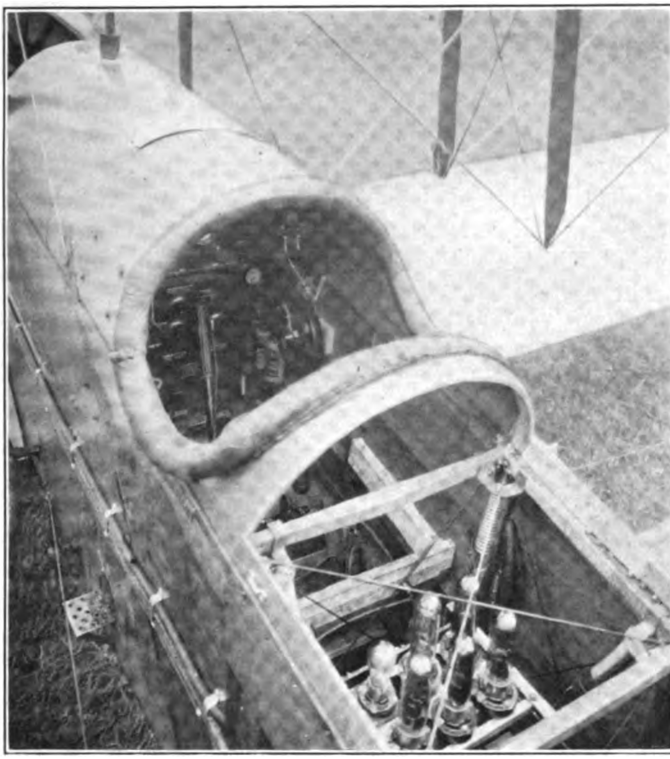


FIGURE 53—Radio Installation of Air Mail Plane, Showing Transmitter Receiver, Reel and Control Box

SELF-RECTIFYING TRANSMITTERS

Two types of transmitters have been produced, utilizing the so-called "self-rectifying circuit," that is, an a. c. plate supply is fed directly to the plates of the oscillating tubes, resulting in a continuous wave output, modulated at a frequency of 1,000 cycles.

The first of these is a "fog signal" transmitter, developed and designed for the United States Department of Commerce, Bureau of Lighthouses, for use on light ships and at shore stations, to enable ships to take radio compass bearings.

The design of this equipment was influenced by it being required that as much as possible of the spark transmitters used at present for this purpose be utilized in the tube transmitters. Also, it was necessary that the circuit used be one of extreme simplicity and reliability.

The circuit used for this transmitter is shown in Figure 54. It utilizes 2 UV-204-A tubes in a self-rectifying Hartley circuit, giving a continuous wave output modulated at 1,000 cycles. A separate machine is used to obtain filament supply, due to the poor regulation of the spark alternator which is used for plate supply, and also to introduce a frequency difference between the filament and plate supplies, thereby insuring longer filament life incident to a. c. filament heating. This practice has been adopted after full consideration was given to voltage regulating equipment which would permit filament illumination from the plate alternator.

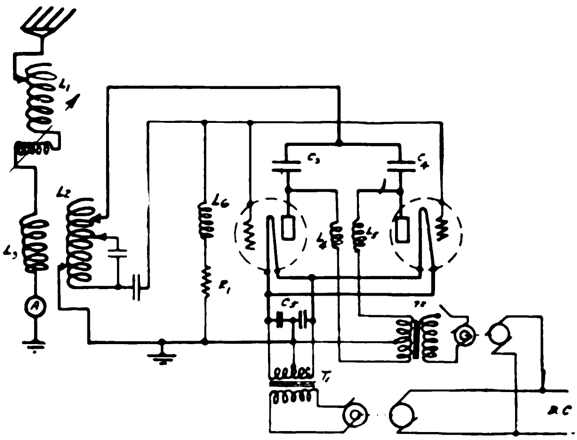


FIGURE 54

An intermediate circuit is used for the suppression of harmonics, since the primary object of the set is to eliminate the interference now being experienced with spark sets.

The keying equipment formerly used with the spark sets is utilized, operating in the primary of the plate transformer. The keying is automatic.

The transmitter is shown in Figures 55 and 56. The power input and power control circuits are brought out on a terminal board mounted back of a door on the lower section of the panel. This terminal board contains the main line contactor and the motor starting contactor.

This set is extremely simple in construction, due primarily to its being built for a very narrow wave length band, and for continuous wave telegraph transmission only. The plate transformer and the converter for filament supply are located at the bottom of the structure as shown in Figure 56. The tubes are mounted on a shock-absorbing mounting and are visible from the front of the panel. A lock is provided for the antenna variometer.

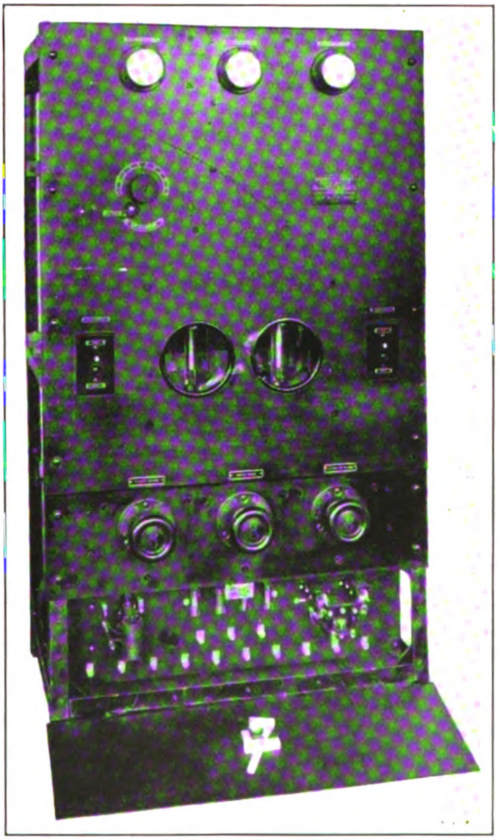


FIGURE 55—Radio Fog Signal Tube Transmitter

A second transmitter, designed to utilize existing spark equipment, is shown in Figure 57, in which the "tube attachment," as it is called, is shown mounted aside of a 2-kilowatt type P-8 Marconi spark transmitter. The design is such that the tube attachment lines up with the spark set.

The circuit for this transmitter is practically identical with

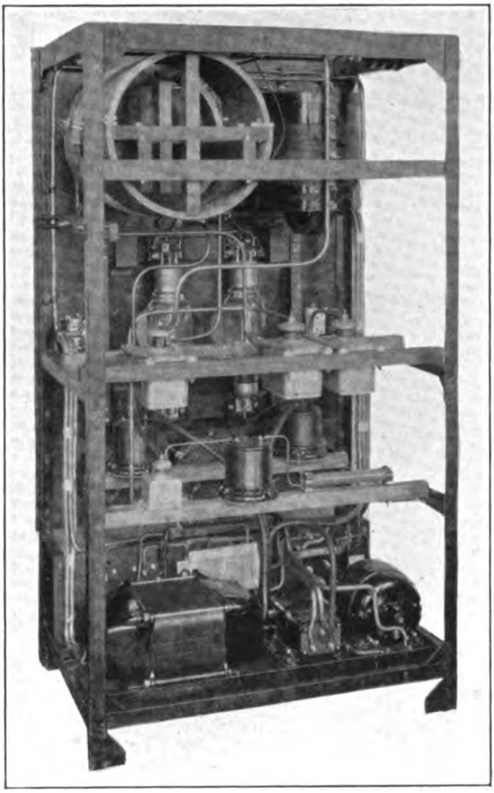


FIGURE 56—Radio Fog Signal Tube Transmitter
(Rear View)

that of the fog signal transmitter just described. It also utilizes two Model UV-204-A tubes, in a self-rectifying circuit, and receives its power from the alternator of the spark transmitter. The set covers a wavelength range of from 2,000 to 2,400 meters. A transfer switch is included in the rear of the set, operated from the front of the panel, which transfers the power supply from the spark to the tube attachment. Plugs are provided at the top of the panel to transfer the antenna from the spark set to the tube attachment, and to select taps on the loading inductance of the latter. Figure 58 shows another view of the tube attachment, indicating its structure more clearly.

BROADCASTING EQUIPMENT

It is not intended to take up a discussion of broadcasting equipment as such, in this paper, altho it is believed that a brief description of the recent equipment installed by the General

Electric Company at Oakland, California, and known as Station KGO, will be of interest.

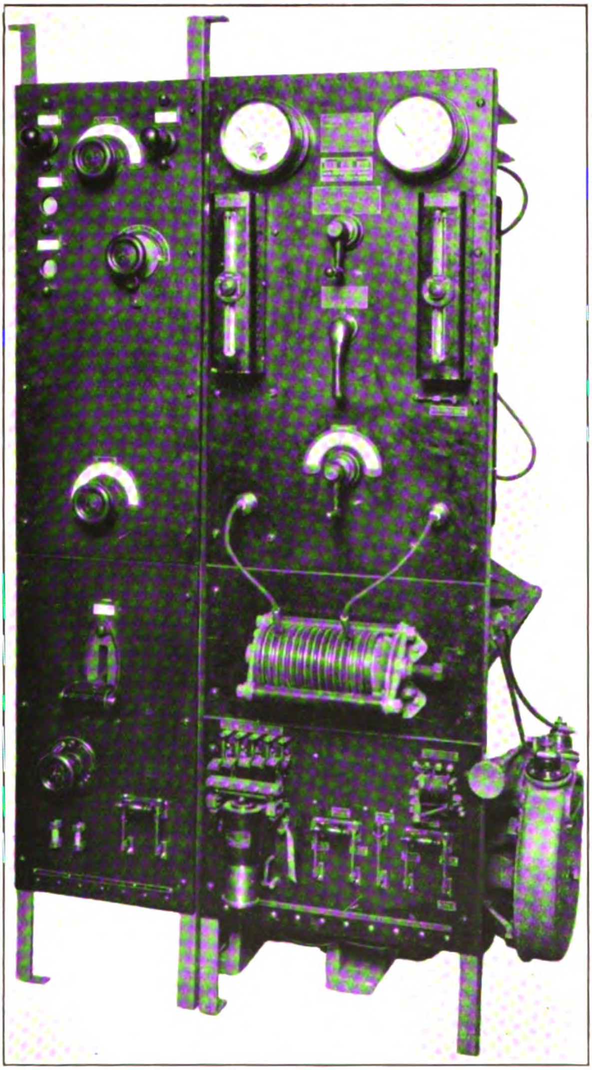


FIGURE 57—Vacuum Tube Transmitter Attachment for Use with Spark Transmitters

The transmitter at KGO operates from a 220-volt, 3-phase, 60-cycle power supply. The main power panel for the station is shown in Figure 59. This panel contains the equipment necessary for the control distribution of power, and the necessary

motors, transformers and other apparatus. The power panel consists of four slate switchboard units. Controls and meters are provided for the excitation of generators, the operation of control circuits, kenotron filament supply, radiotron filament supply, modulator, and power amplifier units. Controls are also provided for the distribution of the power supply to the various amplifiers.

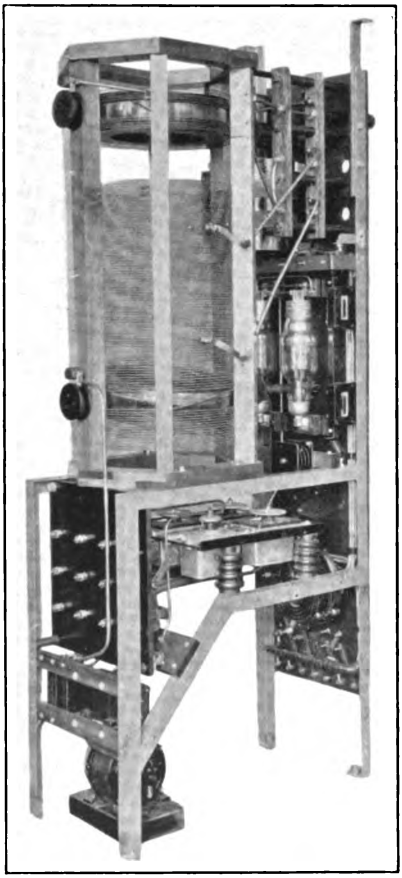


FIGURE 58—Vacuum Tube Transmitter Attachment for Use with Spark Transmetters

Figure 60 shows a rear view of the same panel.

A plate transformer is used to step up the supply voltage to the operating voltage of the kenotron rectifier. The kenotron rectifier, with its associated filtering circuit, is shown in Figure

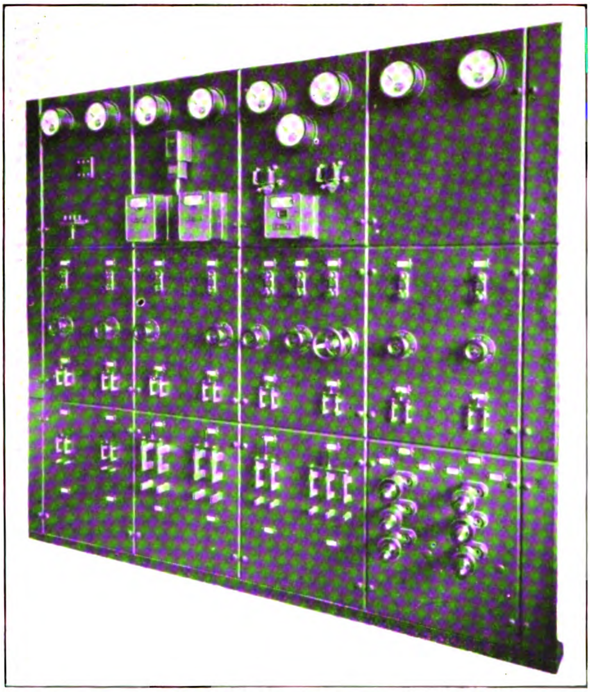


FIGURE 59—Power Control Panels (Front View)

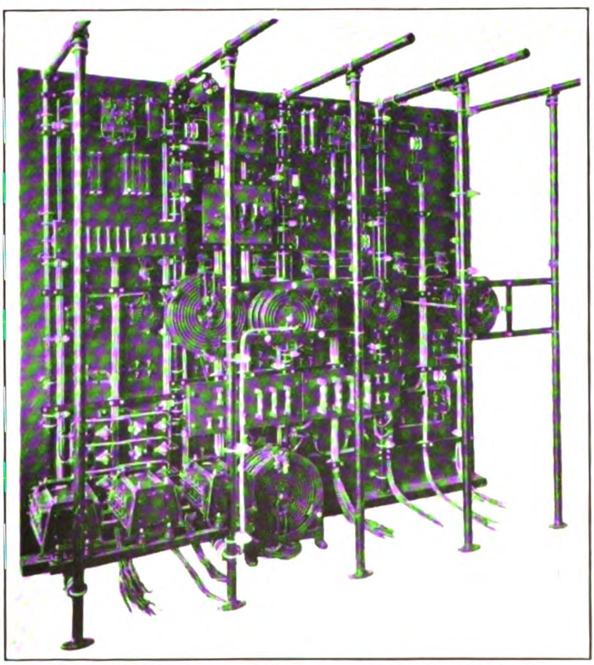


FIGURE 60—Power Control Panels (Rear View)

63. This rectifier provides a 15,000-volt direct current supply, with a ripple of less than one-tenth of one percent. An auto transformer is supplied with the rectifying equipment to provide a variation in output voltage, from 4,000 to 15,000 volts. A rear view of the rectifier is shown in Figure 64. It will be noticed that smoothing condensers are supplied considerably in excess of those required for commercial telegraph or telephone communication.

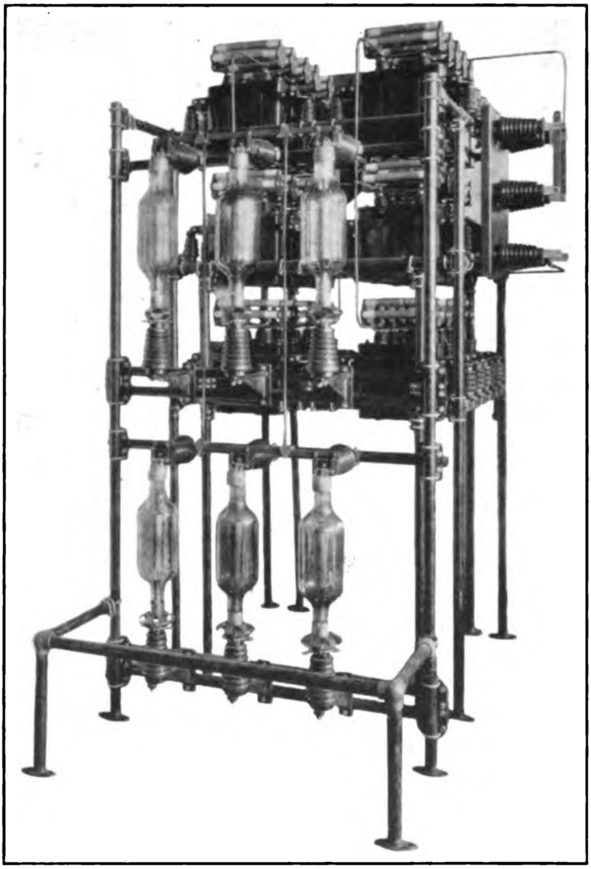


FIGURE 63—Kenotron Rectifier Unit

Figure 65 represents the "grid tuning unit." This unit, as is indicated in the photograph, is carefully shielded, inasmuch as its circuits are depended upon for maintaining constant frequency of output. This unit is designed to permit extremely close adjustment.

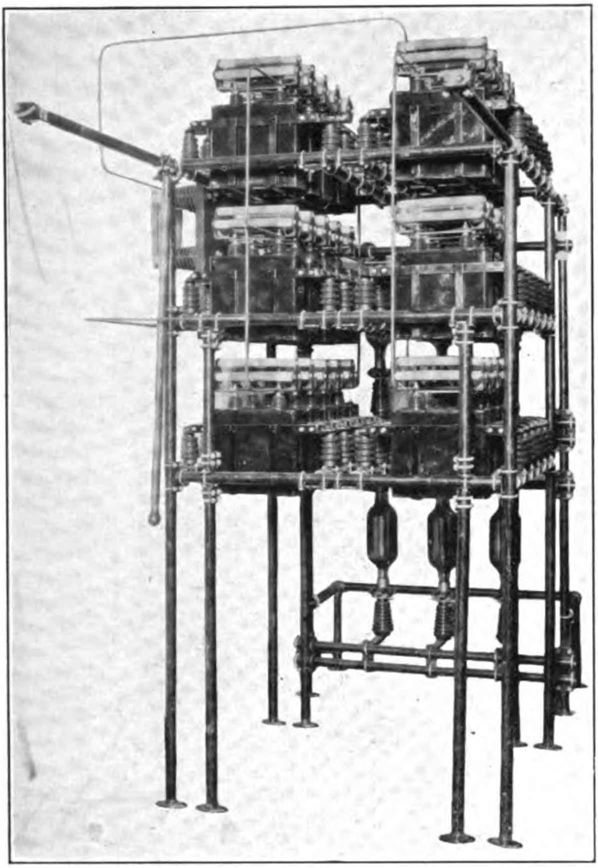


FIGURE 64—Kenotron Rectifier Unit

The oscillator tuning unit is shown in Figure 66. The rear view of this unit is shown in Figure 67. This unit provides an intermediate circuit which eliminates harmonic rediation from the antenna, and in addition, includes the transformer for coupling to the antenna. The edgewise wound coils mounted vertically, indicate the coupling transformer. The coil to the left is pivoted on the periphery, and coupling is varied by a control located on the front of the panel.

The control room equipment consists chiefly of three amplifier and control banks. The first of these, shown in Figure 68, is the 5-watt microphone amplifier bank, containing the various low-powered amplifiers for the pick-up devices used in the studio. Means are provided whereby any pick-up device can be switched to the desired group of amplifiers. A rear view of the 5-watt

amplifier bank is shown in Figure 69. The structure is composed of a number of sub-assemblies, any of which can be readily removed from the main structure for inspection or repair. Each section is completely shielded.

FIGURE 65—Grid Tuning Unit (Front View)

The output of the microphone amplifier bank is carried to the 50-watt intermediate amplifier bank shown in Figure 70. The amplifiers in this bank are arranged so that they may be switched rapidly to prevent any interruption in the program. Figure 71 shows a rear view of this panel. Figure 72 illustrates the monitoring control bank, and among other things, this assembly contains the time signal receiver, radio receivers for checking quality, an antenna power indicator, and the various selector relays. From this bank, the circuit passes underground to the power station. A rear view of this panel is shown in Figure 73.

Batteries are provided for the filament and plate supply of the amplifier tubes, to insure freedom from commutator ripple. A common control panel, illustrated in Figure 74, is supplied for charging all batteries associated with the station.

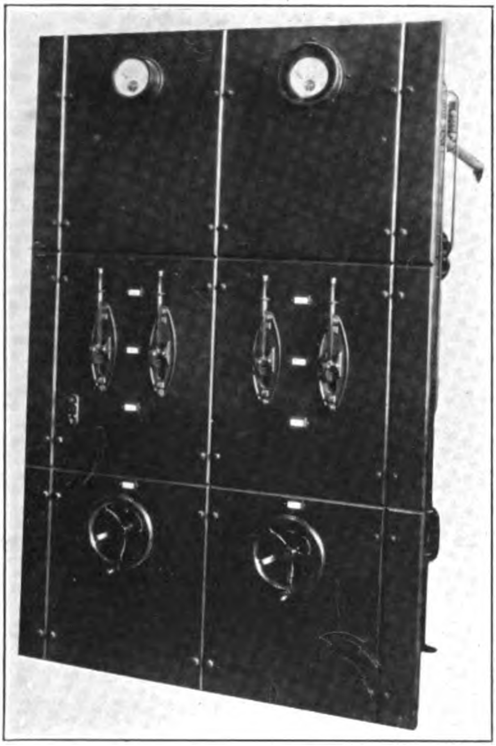


FIGURE 66—Oscillator Tuning Unit (Front View)

It is hoped that a future paper before the INSTITUTE will enter into a complete discussion of the functioning of this broadcasting station. Such discussion, however, is beyond the scope of the present paper. Figure 75 is shown, in closing, to illustrate the vacuum tubes which are utilized in transmitters described in this paper.

The first three of these are receiving tubes. The remainder are transmitting tubes, and are rated at 5, 50, 250, 1,000, 5,000, and 20,000 watts, respectively. All of these tubes, as now manufactured, include thoriated filaments, which has resulted in an appreciable decrease in the power required for the heating of the filament, and an increase in the filament life of the tube.

The 20-kilowatt tubes must be operated in conjunction with a water-cooling system. The anode is exposed directly to the circulating water.

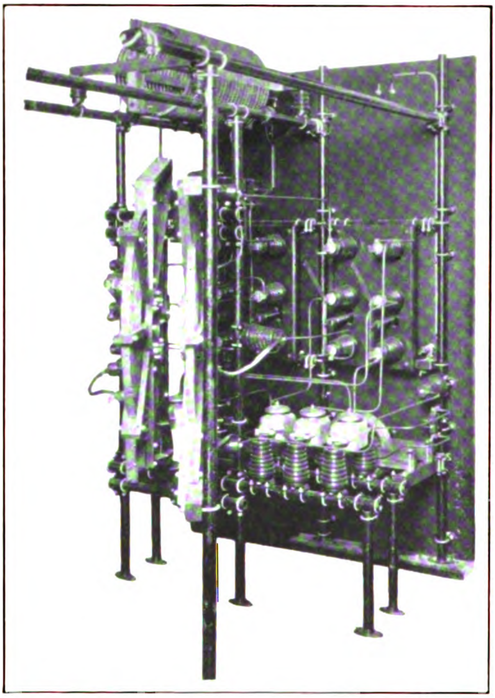


FIGURE 67—Oscillator Tuning Unit (Rear View)

The characteristics of the thoriated filament tube are such that filament voltage regulation becomes a matter of lesser importance. It has been found that the output of transmitters utilizing these tubes is constant thruout a comparatively wide variation in filament voltage, also, that the filaments have sufficient emission so that they can usually be operated at somewhat below rated voltage. It is apparent, therefore, that normal variations in supply voltage will not affect the output of the transmitter, nor decrease the filament life. The greater emission also practically eliminates change in frequency, due to change in filament temperature over a comparatively wide range of filament voltage.

SUMMARY: This paper describes a number of vacuum tube transmitters which have been developed and built during the past year, both for commercial use and the various governmental departments. It includes a discussion of the

circuits utilized, and the reasons for adopting the present circuits in place of the former "antenna oscillators."

The 20-kilowatt transmitters built for the United Fruit Company for installation in Central America, are described in detail, since they represent most completely the developments to date in medium power vacuum tube transmitters.

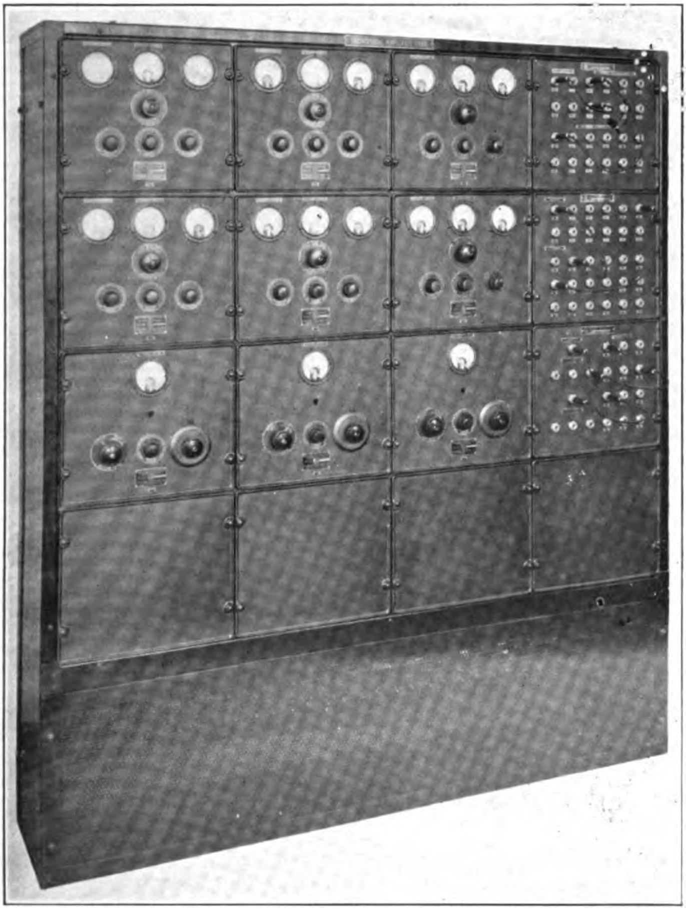


FIGURE 68—5 Watt Amplifier Bank (Microphone Amplifier)

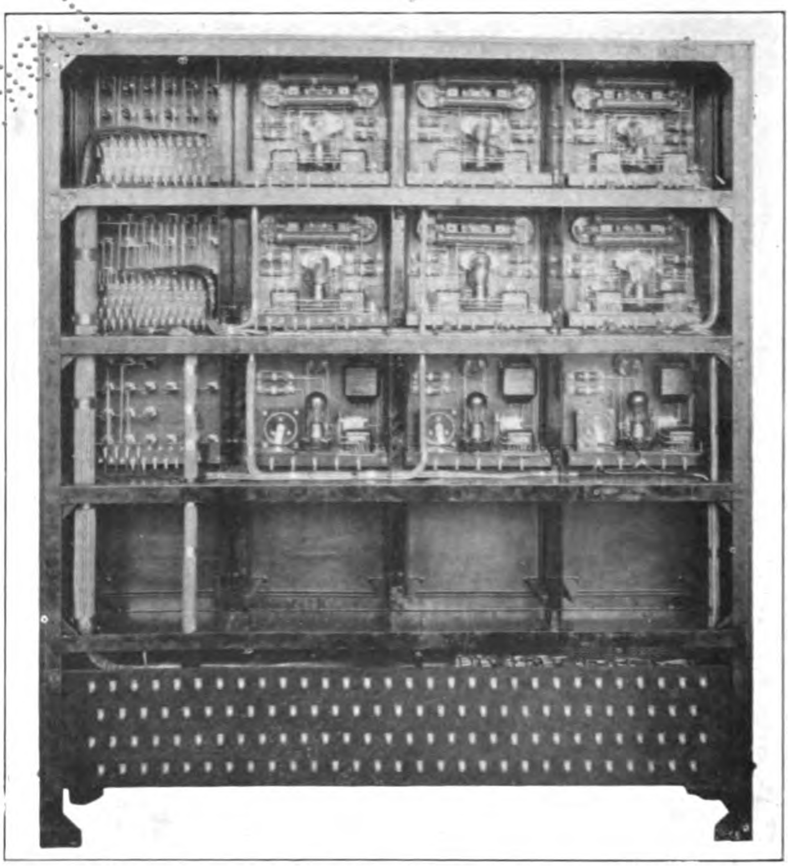


FIGURE 69—5-Watt Amplifier Bank (Microphone Amplifier)

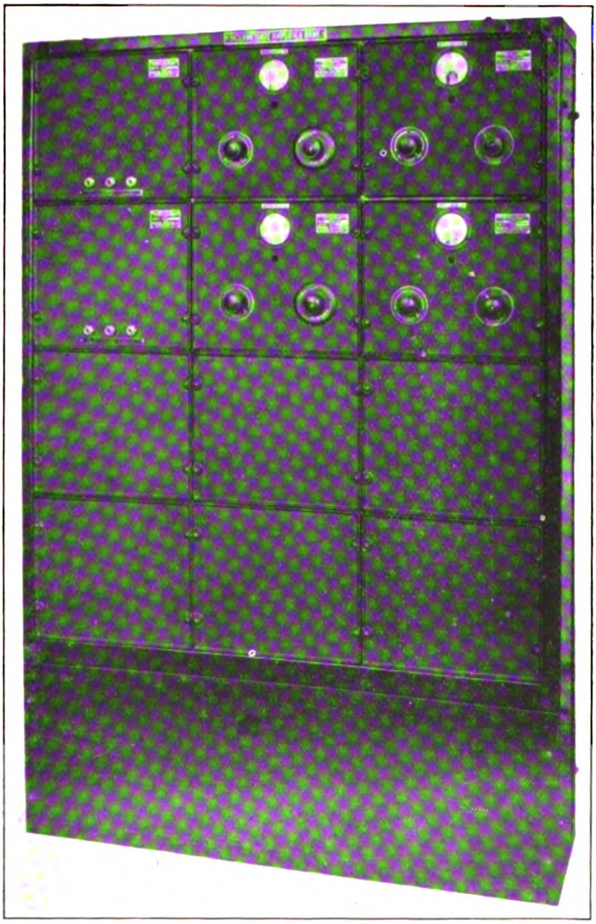


FIGURE 70—50-Watt Amplifier Rack (Intermediate Amplifier Bank) (Front)

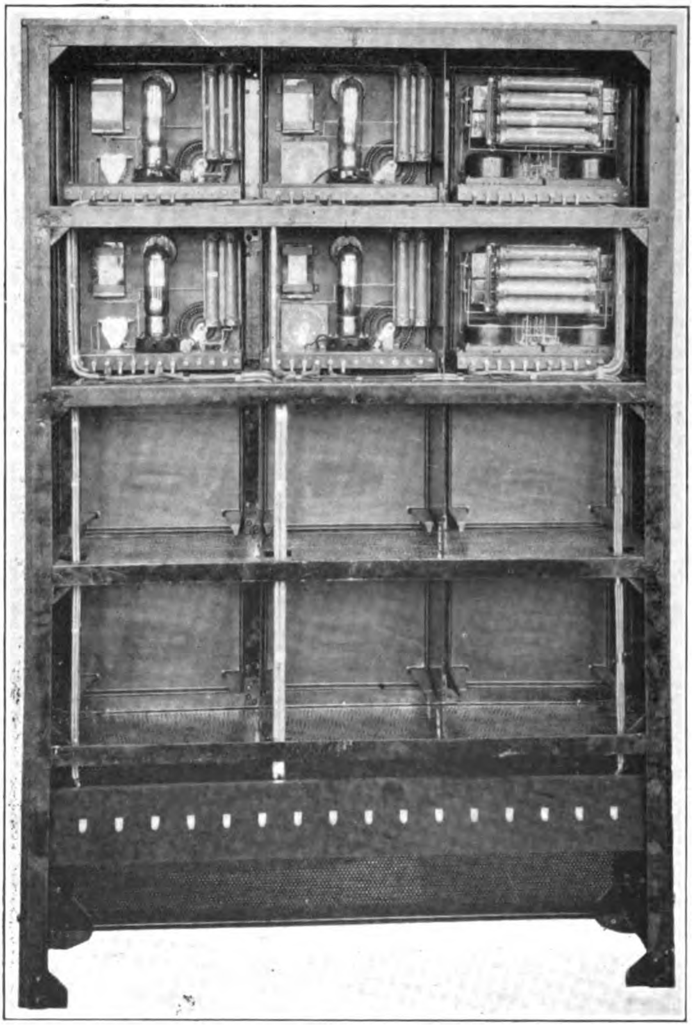


FIGURE 71—50-Watt Amplifier Rack (Intermediate Amplifier Bank) (Rear)

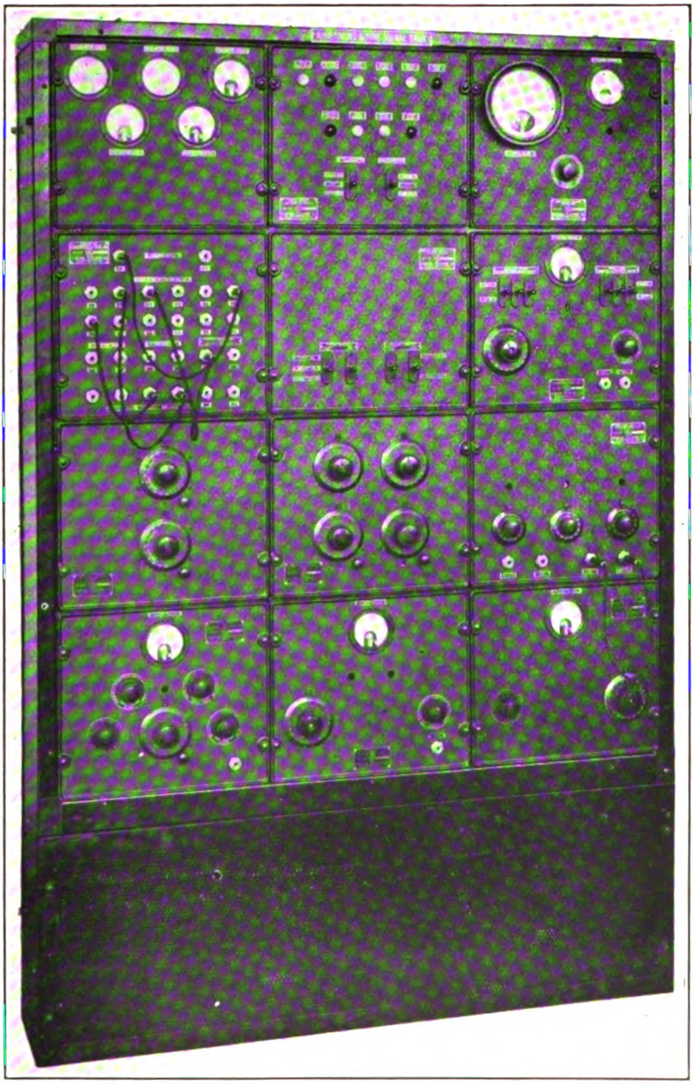


FIGURE 72—50-Watt Amplifier Rack (Monitoring and Time Signal Bank) (Front)

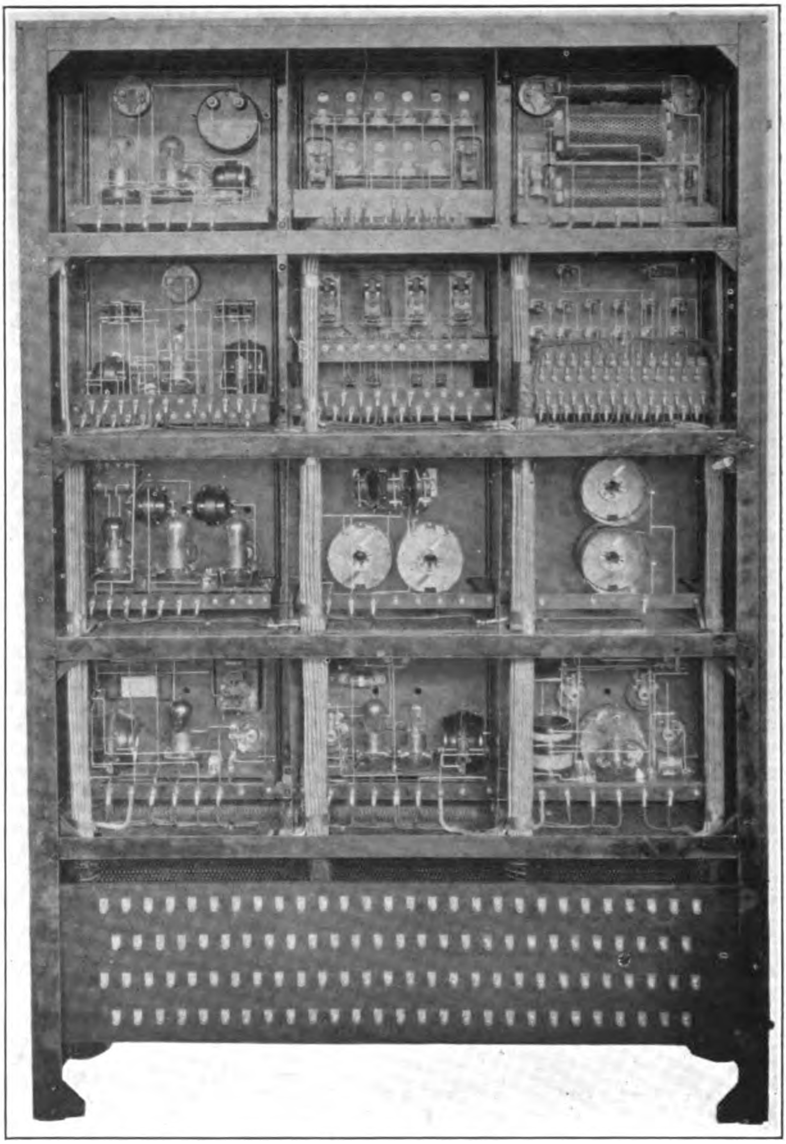


FIGURE 73—Monitoring and Time Signal Bank (Rear)

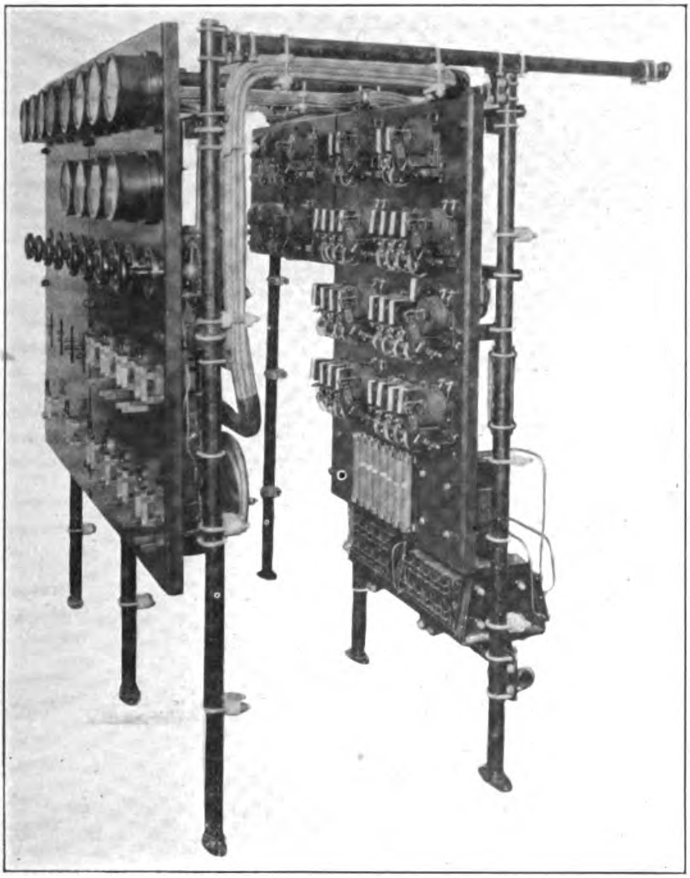


FIGURE 74—Battery and Generator Control Panel (Side View, Front)

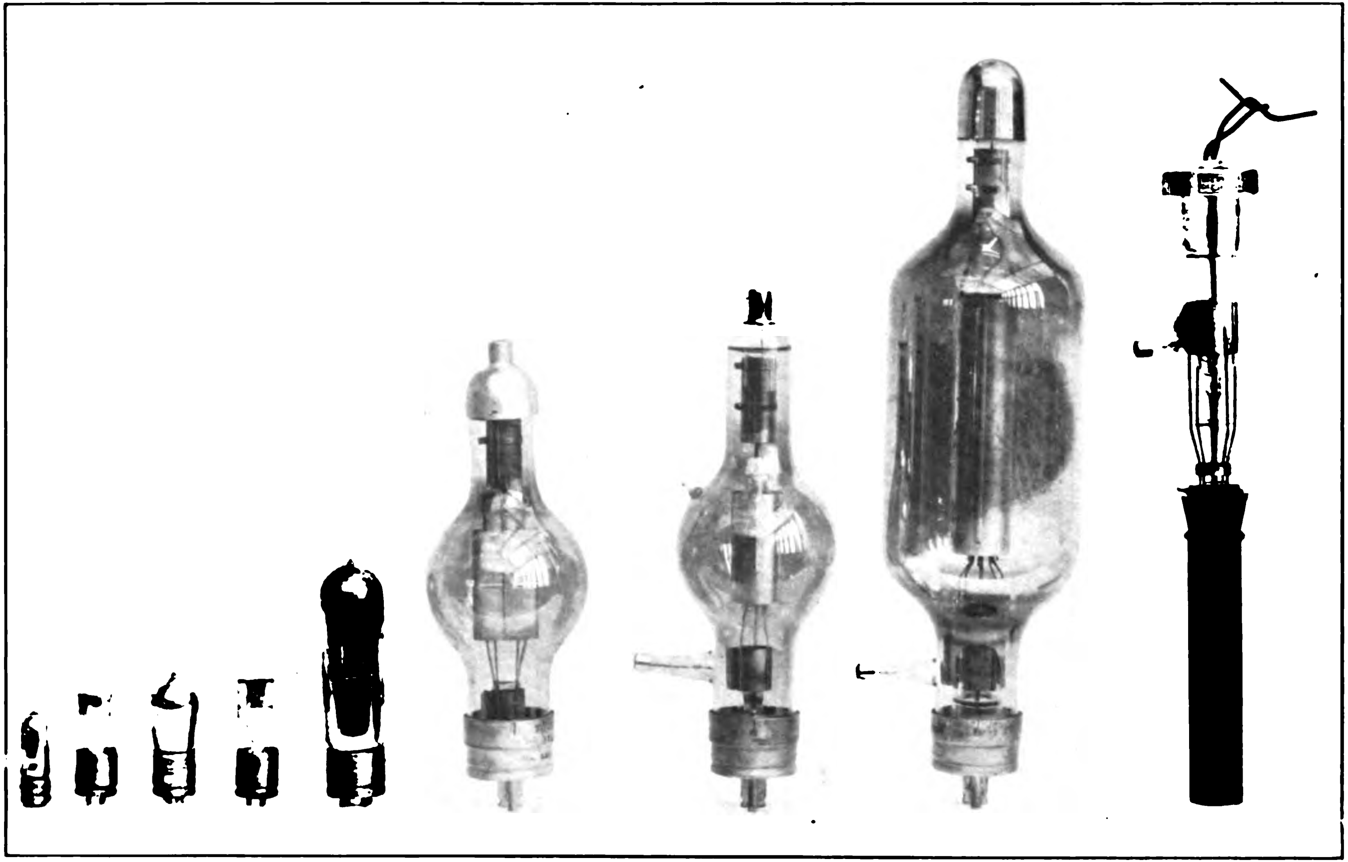


Figure 75—Radiotron Vacuum Tubes

A METHOD OF MEASURING AT RADIO FREQUENCIES THE EQUIVALENT SERIES RESISTANCE OF CONDENSERS INTENDED FOR USE IN RADIO RECEIVING CIRCUITS*

By

Charles N. Weyl

(Moore School of Electrical Engineering)

and

Sylvan Harris

(Consulting Engineer)

Introduction

The purpose of this paper is to show the desirability of measing at radio frequencies the equivalent series resistance of condensers intended for use in radio receiving circuits, and to set forth a method, recently developed by the writers of this paper, of making such measurements. It is believed that the method to be described will serve not only as an aid to the choice of particular condensers for particular purposes, but also that its use will lead to a better understanding of the theoretical and practical aspects of condenser design.

No detailed theoretical discussion of the factors which enter into the losses of condensers will be considered. For such material as is available on this subject the reader is referred to the scientific papers which have been issued from time to time by the United States Bureau of Standards.[1]

Practically every radio receiving circuit makes use of the phenomenon of resonance. In a vast majority of cases it is desirable to produce as sharp a resonance as possible, which indirectly means that all losses of power in the circuit must be reduced to a minimum. Since a resonant circuit consists essentially of an inductance (usually a coil of some sort) and a capacity (some type of condenser) it is important that the designer reduce the losses in both coil and condenser as far as is possible.

Much work has been done on the resistance of coils used in

*Received by the Editor, September 6, 1924.

[1] See particularly Article 34, Bulletin 74. Bureau of Standards.

radio frequency circuits, but the writers of this paper have found that it is almost universal practice to determine the losses of receiving air condensers at audio frequencies, even tho the condensers are intended to be used in radio frequency circuits. This means that it has been thought possible, from the equivalent series resistance measured at audio frequency (for example 1 kilocycle) to compute the value for radio frequencies (for example 1,000 kilocycles). If the results of such a transformation are to be of value, the exact law connecting loss and frequency must be known, since the orders of magnitude of audio and radio frequencies are so vastly different.

A study of the literature has led the writers to believe that these laws are not accurately known and at best would not yield readily to computation. We quote as follows: "The variation of the power factor and the equivalent resistance with frequency is a complicated matter, the laws of which are not accurately known. To a first approximation, however, the power factor of an absorbing condenser is constant. Since $r\omega C$ (power factor) is approximately constant, r (equivalent series resistance) is inversely proportional to frequency."[2] The curve which accompanies the material just quoted shows the variations of the equivalent series resistance of a glass plate condenser between 750 kilocycles and 97 kilocycles. This curve shows *definitely* that the inverse proportionality is not strictly true. Certainly extrapolation from 1 kilocycle to 1,000 kilocycles would be entirely unjustified.

A brief consideration of the elements which constitute the equivalent series resistance of a condenser will serve to show a few of the difficulties involved. Leakage resistance, dielectric absorption, and the resistance of the plates and connections all contribute to the losses in a condenser. All of these factors are subject to variation with frequency. It would be necessary to know precisely how each of these factors varies with frequency if we desire to compute the effect at one frequency when measurements have been made at another, bearing a ratio to the first 1-to-1000.

We shall stop to consider only one of these factors. At 1 kilocycle the skin effect in a parallel plate condenser is small, in fact, practically negligible. At 1,000 kilocycles the skin effect is unquestionably no longer negligible nor is it readily calculable to any reasonable degree of accuracy.

In recent months many laboratory tests have been made by various manufacturers to determine the equivalent series resist-

[2] Bulletin 74, Bureau of Standards, page 125, and following.

ance of condensers. These results have either been stated to refer to 1 kilocycle, which of course is not a useful figure for the radio engineer, or the result obtained from a 1-kilocycle test has been referred to some radio frequency assuming that the equivalent series resistance was *exactly* inversely proportional to the frequency. Since this assumption is not true, such radio-frequency values are practically worthless.

It was with this in mind that the writers developed the method of measurement at radio frequencies of the equivalent series resistance of condensers intended for use in receiving circuits.

The Principle of the Method of Measurement

The principle upon which the method is based can be explained by reference to Figure 1. The coil O is supplied with energy at any desired radio frequency by a vacuum tube oscillator. The loop L and the condenser C form a simple resonant circuit. A known resistance r_s is short-circuited by a heavy removable bar S. A is a thermo-galvanometer.

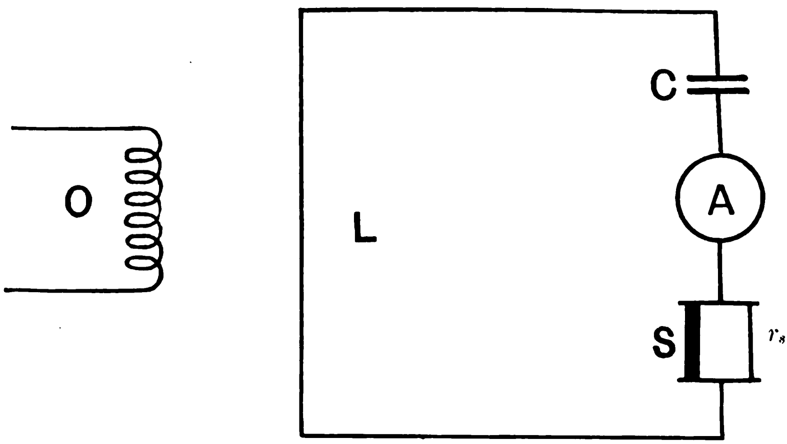


Figure 1

Suppose that the oscillator is adjusted to give resonance in the circuit $L\ C\ A\ S$.

If E is the emf. induced in the circuit by O,

r is the resistance of the loop and ammeter,

r_c is the equivalent series resistance of the condenser.

then $I_1 = \dfrac{E}{r+r_c}$ where I_1 is the resonant current.

Now if the bar S is removed and resonance still obtains

$$I_2 = \frac{E}{r + r_c + r_s}$$ where I_2 is the resonant current.

The justification for assuming E the same in both cases will be made later.

From the above equations

$$r_c = \frac{r_s}{\frac{I_1}{I_2} - 1} - r \tag{1}$$

If r, r_s, and $\frac{I_1}{I_2}$ can be accurately determined we have a means of determining r_c.

It is at once apparent that unless r is smaller than r_c, a small error in the determination of the value of r will introduce a large error in the computed value of r_c. Therefore the resistance of the loop and meter must be as small as possible. The ratio $\frac{I_1}{I_2}$ must also be as large as is consistent with the accurate part of the range of the galvanometer being employed. Further, if the equivalent series resistances of condensers of different capacities are to be measured at the same frequency, a loop of variable inductance must be used.

Preliminary experiments showed that for air condensers having a capacity of the order of magnitude of 500 micro-micro farads, the equivalent series resistance was in the neighborhood of an ohm at 1,000 kilocycles. Therefore it was necessary to make the loop and galvanometer resistance as much less than this as was possible.

A very sensitive thermo-galvanometer having a resistance of about 5 ohms was chosen. This meant, of course, that it was necessary to shunt the galvanometer with a low resistance.

However, reference to article 42, Bulletin 74, Bureau of Standards explains the difficulty of employing such shunts due mainly to the impossibility of determining accurately the division of the current between the shunt and the meter. It is to be noted that this difficulty was eliminated in the method used by the writers, since the expression for r_c depends upon the *ratio* of I_1 to I_2 and not upon their absolute values. It was assumed, and later justified, that altho the distribution of the current between shunt and meter could not be accurately determined, that *for a given frequency* the ratio of shunt current and meter current would be the same for different values of total current thru the loop.

Since a thermo-galvanometer was used, *for a given frequency*, the readings of this meter, which had been carefully checked

against a standard, were *proportional* to the square of the current thru the loop and condenser.

As a shunt for the galvanometer, a three-inch (7.6 cm.) length of number 14 copper wire was used, which even at radio frequencies had a negligible resistance.

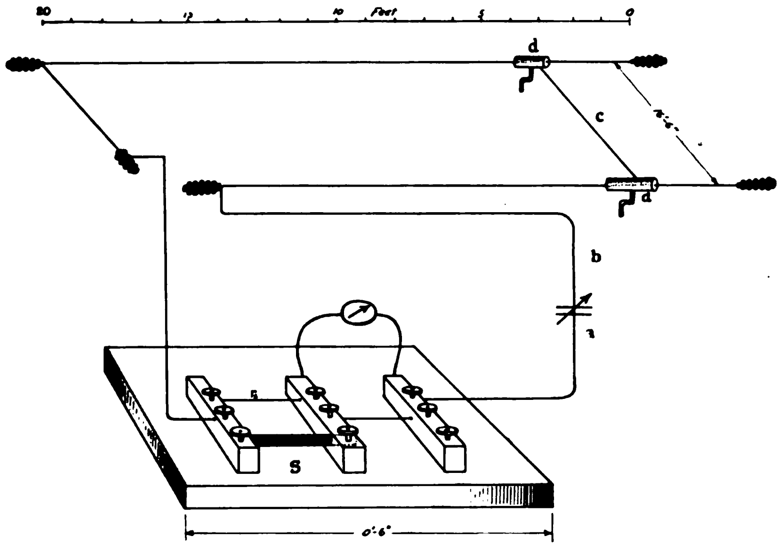


FIGURE 2

Before building the loop, curves of inductance and resistance per foot for various sizes of copper wire were calculated from formulas given in a paper by Rosa and Grover, Bulletin, Bureau of Standards, volume 8, 1912. The results of these computations appear in Figures 3 and 4.

The adjustable loop finally chosen (Figure 2) was made of number 8 Brown and Sharpe, copper wire and had maximum dimensions of 17.5 by 25 feet. The resistance of this loop for various adjustments is shown in Figure 5. With this loop, resonance was obtainable at 1,500 kilocycles for the capacities ranging from 250 to 500 micro-microfarads. At this frequency the total resistance of the loop including leads was calculated to be 0.8 ohms. No advantage could be gained by using heavier gauge wire since the increased diameter of the wire would necessitate the use of a larger loop, because the inductance of a loop of given dimensions decreases as the diameter of the wire increases.

However since the resistance of such a loop can be calculated with great accuracy and since the equivalent series resistance

of a 250 micro-microfarad condenser is approximately 2 ohms at 1,500 kilocycles, this loop was quite satisfactory.

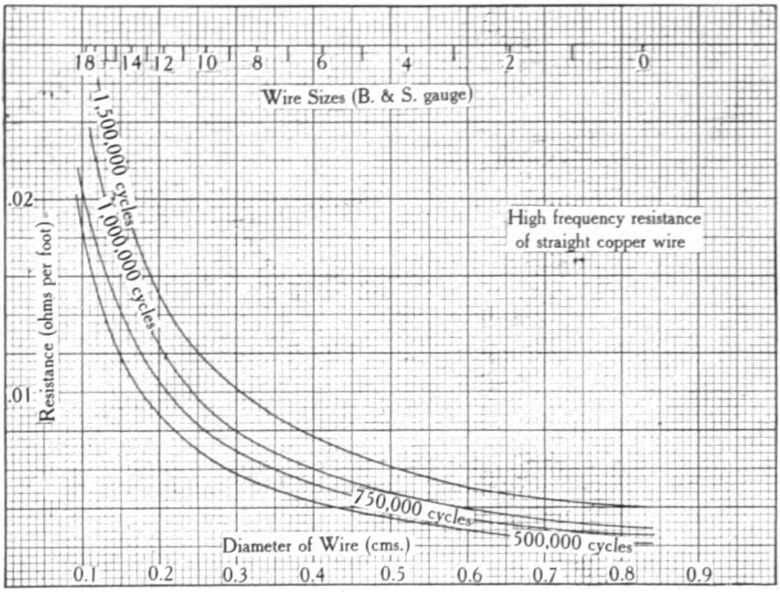


FIGURE 3

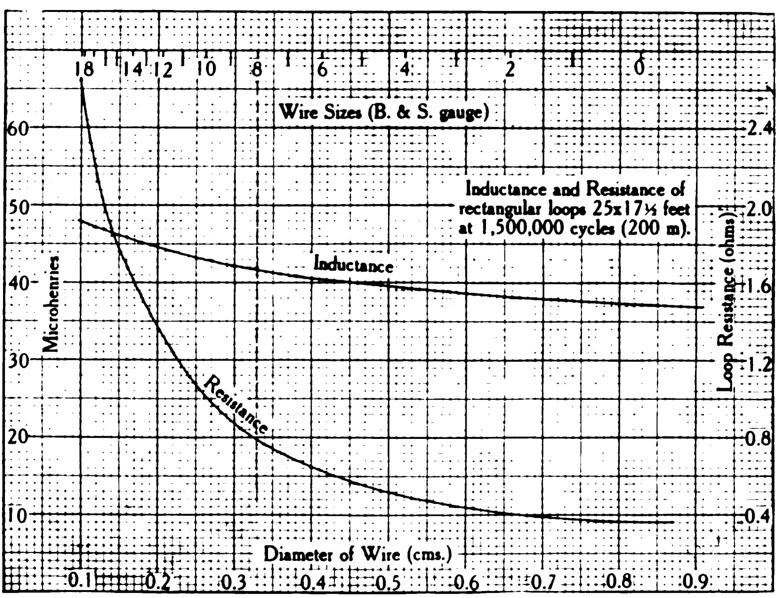


FIGURE 4

It is to be noted that larger condensers which, in general, have smaller resistances, required less loop and consequently r was proportionately decreased.

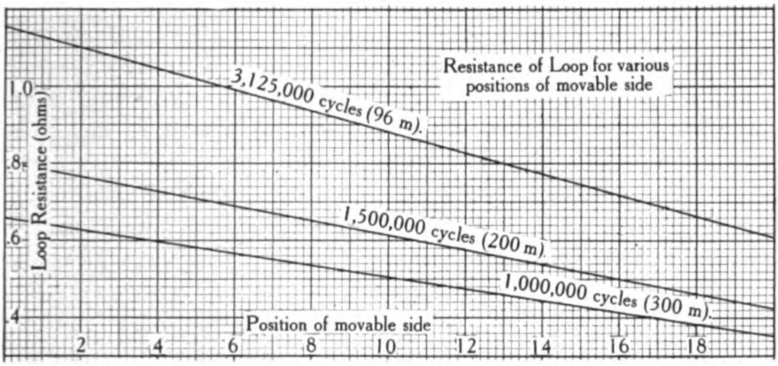


FIGURE 5

The resistance r_s (Figure 2) was a single strand number 30 Brown and Sharpe, manganin wire having a d-c. resistance (measured in place, by a Wheatstone bridge) of 2.10 ohms. Reference to Table 1, Chapter 2, Morecroft's "Radio Communication," will show that the resistance of a manganin wire of less than 0.29 mm. in diameter at 3,000 kilocycles or less, differs by less than one percent from the d-c., resistance. Manganin was chosen on account of its very low temperature coefficient of resistance.

The short-circuiting bar was made of copper and could be clamped into position by heavy set screws (Figure 2). It is important to note that the short-circuiting bar and the resistance r_s, were offset with respect to the loop in such a manner as to make the inductance of the loop virtually the same with the bar in or out.

A twenty-watt oscillator was employed with extremely loose-coupling to the loop. In this way a negligible reaction took place between the loop and oscillator.

PROCEDURE OF MEASUREMENT[3]

A measurement is made in the following manner:

1. The condenser is connected to the points a and b (Figure 2) with short heavy leads.

[3] All experimental work was conducted in the laboratories of the Moore School of Electrical Engineering, University of Pennsylvania.

2. The dial is set at the desired capacity, if a variable condenser is being measured.

3. By means of an accurate wavemeter the oscillator is adjusted to give the desired frequency.

4. With r_s short-circuited by means of bar S the adjustable side, c, of the loop is moved until resonance is indicated by the thermo-galvanometer. Non-conducting rods should be used to make the final adjustment of the loop in order to avoid body capacity effects. When the exact point of resonance has been found the clamps, d, are tightened.

5. The output of the oscillator is then adjusted to give approximately full scale reading of the thermo-galvanometer, the frequency being maintained constant, using the wave meter as a check.

6. A reading is taken of the thermo-galvanometer, the observer being careful not to touch any part of the apparatus.

7. The short-circuiting bar S is then unclamped and slid back, introducing the additional resistance r_s into the circuit.

8. A second reading of the thermo-galvanometer is taken.

9. The position of the adjustable side of the loop is noted.

Calculation of the Equivalent Series Resistance

Having obtained the ratio $\frac{I_1}{I_2}$ and the position of the adjustable side of the loop, the calculation of the equivalent series resistance is quite simple. By consulting the curves in Figure 5, the resistance of the loop and leads are determined at the given frequency. These data are then substituted in equation (1), which yields r_c directly.

Proof of the Validity of the Method

The equivalent series resistance of a 500 micro-microfarad variable air condenser was measured at 1,500 kilocycles by the method described above. A piece of number 30 Brown and Sharpe "Advance" wire was then carefully soldered between two heavy copper lugs. A careful measurement by means of a Wheatstone bridge yielded a resistance of 0.775 ohms for this wire. The wire with its lugs was then connected in series with the 500 micro-microfarad condenser and the combined resistance of the condenser and resistance wire was measured at 1,500 kilocycles. The value of r_c was then deducted from the combined resistance of r_c and the "Advance" wire. This yielded a result of 0.796 ohm or a difference of less than 2.7 percent, part of which difference

is accounted for by skin effect. This test was repeated for several known resistances and at several frequencies, the maximum deviation noted being less than 3 percent.

A series of tests were also made to determine the effect of different loop and oscillator couplings and also different values of r_s varying from 0.56 ohms to 2.12 ohms. The deviation from the mean of the values was less than 3 percent.

The effect of foreign metallic bodies in the region of the apparatus was also determined, showing that under the conditions of these tests such effects were entirely negligible.

A careful test was made to determine the effects of any harmonics which might be present in the oscillator output, but the sharpness of the measuring circuit was so great as to render such effects entirely negligible.

Some Results Obtained by the Method

While it is the primary intention of the writers to present the method of measurement in such a fashion that other experimenters may be able to avail themselves of it, for the investigation of condenser losses, it may not be out of place to present briefly some results obtained by means of this method. No attempt whatsoever will be made in this paper to interpret these results.

The curve in Figure 6 is practically self-explanatory, showing the variations of the equivalent series resistance with dial setting of a standard make of air condenser of the so-called "low loss" type. The frequency was constant at 3,125 kilocycles (approximately 100 meters) thruout the test. The large resistance of such a condenser in the low part of its range, is clearly demonstrated.

The curve in Figure 7 shows the variation of the equivalent series resistance of the same condenser with frequency, maximum intermesh of the plates being maintained thruout the test. The regularity of the points is further evidence for the precision of the method. Figure 8 shows the same data plotted on logarithmic paper. While the frequency range is small, this curve indicates that the exponent of the resistance—frequency relation for this condenser is approximately 0.76. A measurement by the Bureau of Standards of this same condenser at 1 kilocycle yielded a result of 278 ohms which justifies an exponent of 0.7.

These results should serve to show the fallacy in referring 1-kilocycle measurements to radio frequencies, using the reciprocal law.

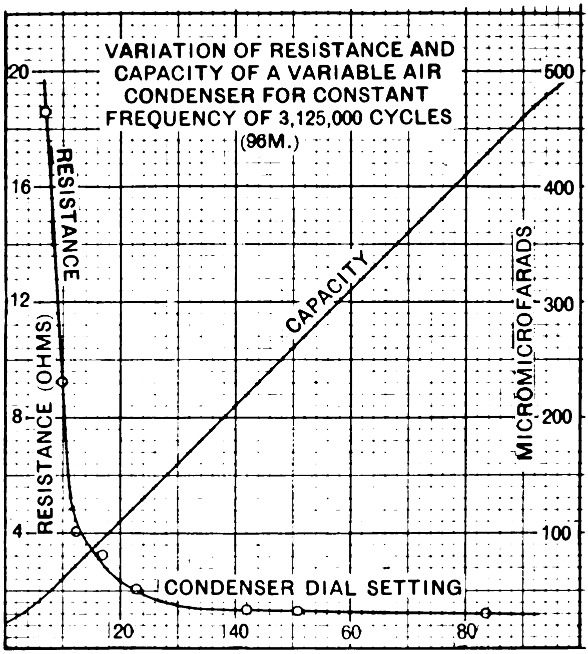


FIGURE 6

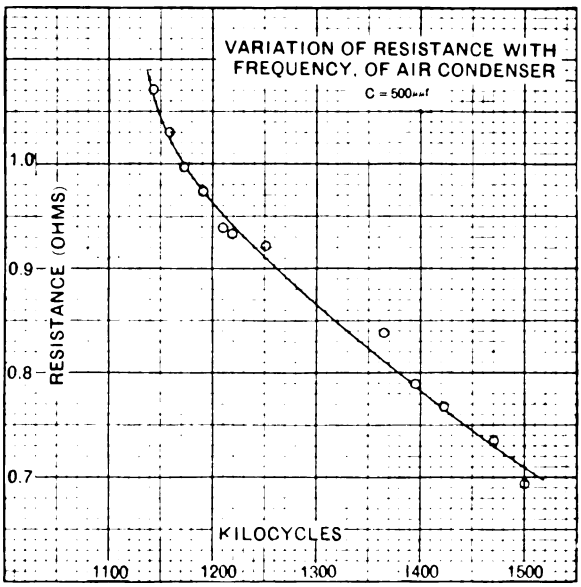


FIGURE 7

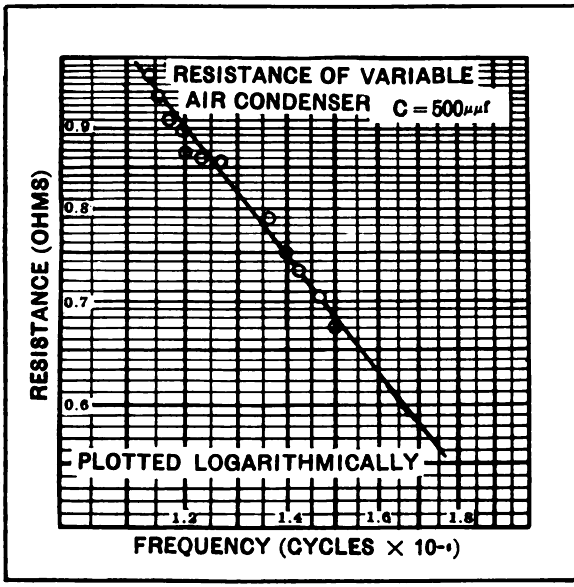


FIGURE 8

Table 1 shows the averages of tests made on over a hundred different air condensers of commercial manufacture.

TABLE 1

AVERAGE EQUIVALENT SERIES RESISTANCE OF AIR CONDENSERS AT MAXIMUM SETTINGS

Capacity of Smallest Cond.	Capacity of Largest Cond.	Average Equivalent Series Resistance	Number of Cond. in group
Micro-microfarads		Ohms at 1,500 Kilocycles	
200	500	1.02	75
500	800	0.75	22
800	2000	0.45	30

Table 2 gives the resistances of four sizes of typical "low loss" condensers at maximum capacity, all of the same make and design.

TABLE 2

EQUIVALENT SERIES RESISTANCES OF A SET OF "LOW LOSS" AIR AT MAXIMUM SETTING, OF SAME CONSTRUCTION AND DESIGN

Capacity in Micro-microfarads	r_c at 1,500 kilocycles
200	1.82
380	0.96
500	0.68
980	0.36

Table 3 gives the resistance at maximum capacity for a set of six condensers showing the effect of enclosing condensers in metal cases.

TABLE 3

SHOWING EFFECT OF METAL CONTAINERS ON AIR CONDENSER LOSSES

Approx. capacity in micro-micro-farads	r_e in ohms at 1,500 kc. with metal container	r_c in ohms at 1,500 kc. without metal container
270	2.02	1.53
515	1.00	0.79
1075	0.57	0.45

A condenser of standard make having solid bakelite end plates was measured at 1,000 kilocycles. The end plates were then drilled leaving only a supporting skeleton of bakelite. A new measurement was made. A pigtail was then added. A third measurement of resistance was made, the results of these measurements are shown in Table 4.

TABLE 4

EFFECT OF SOLID END-PLATES AND PIGTAIL ON LOSSES

800 Micro-microfarad Condenser

Condition	r_c in ohms at 1,000 kilocycles
Solid end plates (no pigtail)	0.85
Skeleton end plates (no pigtail)	0.81
Skeleton end plates and pigtail	0.73

Table 5 is self-explanatory.

TABLE 5

OLD STYLE VERSUS "LOW LOSS" AIR CONDENSERS

Old Style			"Low Loss"		
Average Capacity in micro-micro-farads	r_c ohms at 1,500 kc.	Number Tested	Average Capacity in micro-micro-farads	r_c ohms at 1,500 kc.	Number Tested
500	0.97	33	500	0.73	23
1000	0.57	23	1000	0.41	9

CONCLUSION

The above data and curves are indications of what may be accomplished with this method. The writers hope that its application to the problem of condenser selection and design will yield valuable results in the hands of other investigators as well as in their own.

SUMMARY: This paper shows the desirability of measuring at radio frequencies rather than at audio frequencies, the equivalent series resistance of radio receiving condensers. It further sets forth in detail a method of making such radio frequency measurements, and shows some preliminary results obtained from the application of this method.

THE "PIONEER BROADCASTER"

(Discussions on Mr. D. G. Little's paper on "KDKA, the Radio Telephone Broadcasting Station of the Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania," by Messrs. Howard E. Campbell, D. G. Little, and Lee de Forest; summarized at the request of the Board of Direction of The Institute by Professor John H. Morecroft.)

In an art like radio, which had such a phenomenal growth all over the country, any one's claim to the distinction of being the "pioneer broadcaster" is sure to raise an argument. Mr. Little's designation of KDKA as the country's "pioneer broadcasting station" will not be questioned by most radio readers, but should any one dispute the claim, we must at once admit that the first radiophone broadcasting was not done by the Westinghouse Company at KDKA, but by Lee de Forest, who undoubtedly antedates all other claimants to the title "pioneer."

Ten or more years before any of the others, de Forest threw the human voice over a country almost devoid of listeners; we can well remember listening to him on our crystal receiver sets, and a real thrill it gave. In 1916, radiophone experimenting was carried on at the Highbridge laboratory of the de Forest Company; Governmental regulation stopped broadcasting for a while; then in 1919-1920 we had a real and permanent revival of the art. From the laboratory of Mr. Robert Gowan, in Ossining, the human voice was thrown over the ether waves at this time and, in late 1919, Mr. Frank Conrad, engineer of the Westinghouse Company, began to send out radiophone "programs" from his home. These early phonograph reproduction programs were undoubtedly the private enterprise of Mr. Conrad, and not a Westinghouse activity at all, but Mr. Conrad's success resulted in the transfer of the station from his home to the Westinghouse plant. Thus station KDKA began to function, getting its call letters from the Government on November 5, 1920.

In the meantime a de Forest radio set had been put into operation by the Detroit *News*, and on August 31, 1920, this station sent out election returns. The station was not called WWJ until a year and one-half later, but it was evidently operating from the office of the Detroit *News*, as a news service,

before the activities of Mr. Conrad had been transferred to the Westinghouse factory. So the Detroit *News* station antedates the Westinghouse factory station, the Conrad home station antedates the Detroit *News*, the Gowan home station antedates the Conrad, and de Forest's activities antedate them all. As to which was the "pioneer broadcaster" the reader may judge for himself.

DISCUSSION ON "A METHOD OF MEASURING VERY SHORT RADIO WAVE LENGTHS AND THEIR USE IN FREQUCNCY STANDARDIZATION"

BY FRANCIS W. DUNMORE AND FRANCIS H. ENGEL

BY

EIJIRO TAKAGISHI

AND

SHIGEYOSHI KAWAZOE

(THE ELECTRO-TECHNICAL LABORATORY, MINISTRY OF COMMUNICATIONS, JAPAN)

The above paper on frequency standardization (PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 11, number 5, October, 1923, page 407) interested us greatly, but the writers desire to point out some doubtful points in the paper. The paper states that the authors measured the length of the standing wave produced on parallel wires by marking on the wires the points where a thermo-galvanometer showed a maximum indication of current. According to the observations in our experiments on the same problem it was found that generally there exist two points corresponding to maximum indications of the galvanometer bridged across the parallel wires and between them a point of minimum indication as shown in Figure 1. At either point of maximum indication of the galvanometer, the whole nature of the standing wave on the wires has been so much affected by the insertion of the galvanometer that the ammeter at the point of excitation (or at the current loop) showed a rapid fall of current, but when the galvanometer is put at the point of minimum indication, hardly any decrease in the current has been observed.

From these phenomena they arrived at a conclusion that for the measurement of the wave length on the wires the point of minimum indication should be preferred to those of maximum indication, and this enabled them to get more accurate results. To analyze these phenomena rigorously by the help of mathematics involves many difficulties, but a simple conception may lead us to a fair understanding of the present problem. In Figure

*Received by the Editor, July 14, 1924.

2, the full lines represent the standing wave of current and the dotted ones represent that of potential, the distribution being as usual, and the fall potential due to line resistance being ignored. The potential difference across the both wires will be a minimum at A, A'. Therefore, if an ideal voltmeter which consumes no

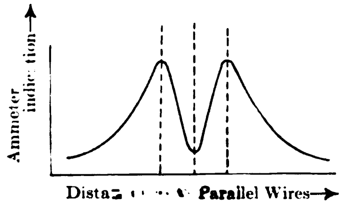


FIGURE 1

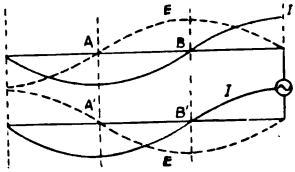


FIGURE 2

current be bridged across the wires, the voltmeter might have a minimum indication at this point, that is, at the anode of potential or the loop of current. If an ideal ammeter which requires no potential drop across it be used instead of the ideal voltmeter, it is clear that it would indicate maximum current at the same point owing to the well-known resonance phenomenon. This being the case it follows that in our practical cases the maximum indication may exist at a certain position other than A, A' and B, B', and if the meter bridged across the wires has a low resistance (as is the case for a usual ammeter) the maximum indication occurs near the points A, $A'$and on both sides of them. In such cases, within a small region very near the minimum point, the resonance condition on the wires is hardly affected by bridging a meter across them, and the indication is determined mainly by the normal form of potential distributions on the wires. It may be deduced also that the greater the resistance of the meter, the further the points of maximum indication may be moved away from the position A, A' and in both directions along the wires, and that if the ammeter has a negligible resistance, the two maximum points may practically coincide with each other. This also was actually observed in our experiments.

F. W. Dunmore and **F. H. Engel** (by letter):* The fact that two current maxima were found by Messrs. Takagishi and Kawazoe in their experiments with the parallel wire method of frequency standardization was no doubt due to the fact that the power output of the generating set was insufficient, thus requiring

*Received by the Editor, August 29, 1924. Published by permission of the Director of the Bureau of Standards of the United States Department of Commerce.

too close coupling between the generating set and the parallel wire system. The effect produced by this combination is quite similar to that encountered in spark transmitting sets when the coupling between the primary and antenna circuits is too tight,—the familiar double hump in the resonance curve is obtained. This "pulling" effect actually changes the frequency of the generating set in the parallel wire method of frequency standardization. The authors completely overcame this effect by using a 50-watt tube as the source of radio-frequency power, and by keeping this generating set as loosely coupled to the parallel wire system as possible. In this way the reaction between the parallel wire system and the generating set was reduced to such an extent that but one sharp hump could be found at each half wave length position along the parallel wire system.

A theoretical discussion of the above point may be found in Bureau of Standards Scientific Paper Number 491, "Theory of Determination of Ultra-Radio Frequencies by Standing Waves on Wires," by A. Hund.

August 15, 1924.
Department of Commerce,
Washington, D. C.

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

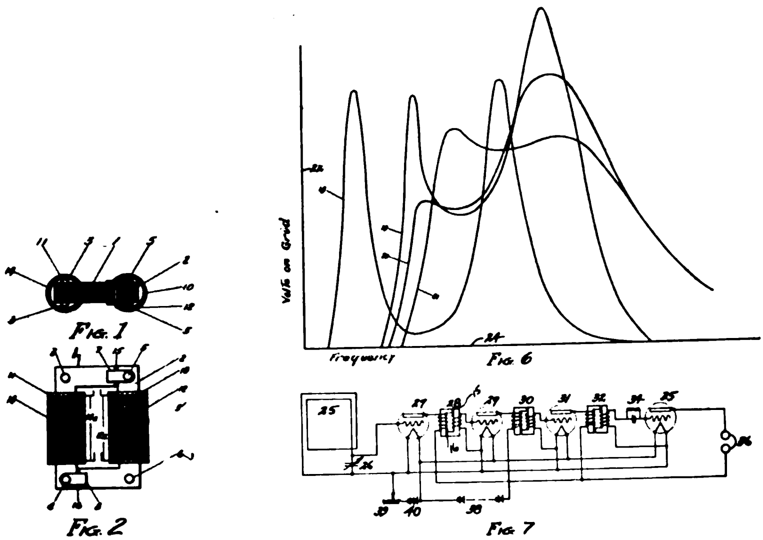
(Issued November 4, 1924–December 30, 1924

By

John B. Brady

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,505,085—C. E. Brigham, filed July 15, 1922, issued August 19, 1924. Assigned to Dubilier Condenser and Radio Corporation, New York.



Number 1,505,085—Radio Receiving Apparatus

Radio Receiving Apparatus, including a radio frequency amplification circuit wherein the tubes are interlinkedby coupling transformers designed to amplify incoming signaling frequencies over a broad band of frequencies. The transformer is designed to have extremely small distributed capacity at the frequencies for which the transformer is designed to operate. A magnetic leakage gap is introduced in the magnetic circuit for

*Received by the Editor, January 10, 1925.

producing losses to a degree necessary to enable the transformer to operate effectively over a broad band of frequencies. The patent shows a number of curves which have been taken on different radio frequency transformers and shows the relatively flat characteristic curve 21 which is possessed by the transformer of the present design.

1,513,973—F. M. Doolittle, filed February 21, 1924, issued November 4, 1924.

RADIO TELEPHONY system of broadcasting wherein the sounds reproduced at the receiving station may be so reproduced as to impart an effect of true tone values derived from a sense of location of the artists or the musical instruments at the broadcast studio. Two radio channels of transmission are provided each under control of separate sound pick-up microphones so relatively positioned as to receive sound in a manner simulating the reception of sounds by the ears of a human being. The radio channels are non-interfering and are each separately adjusted to bring in the transmission from the same studio.

1,514,295—J. F. Lindberg, filed July 5, 1922, issued November 4, 1924.

CONDENSER of the variable type in which both the stationary and the movable plates may be adjusted thru a vernier adjuster. The stationary plates are mounted at peripheries in such manner that they may be rocked with respect to the movable plates while the movable plates may be independently rotated in respect to the stationary plates.

1,514,369—H. A. Bremer, filed September 21, 1923, issued November 4, 1924.

ELECTRIC CONDENSER having a vernier adjustment, consisting of a supplemental plate element engaged by a spring connected with the movable plates and arranged to be rotated with respect to both the movable and the stationary plates. A spring element is included between the supplemental plates and the shaft which carries the movable plates which spring element is in the form of a U-shaped clip.

1,514,648—R. Bown, filed November 12, 1920, issued November 11, 1924. Assigned to American Telephone and Telegraph Company, New York.

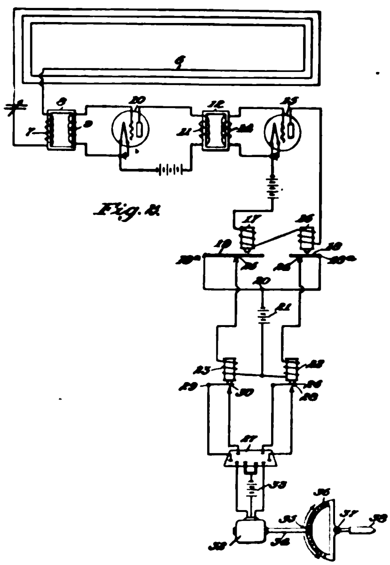
DIRECTIVE RADIO SYSTEM, in which a plurality of antennas

are arranged and each excited by a separate oscillator. The antennas are spaced apart for a distance corresponding to an even multiple of a wave length. A common modulating circuit is provided connecting by line wire with the several stations so that each oscillation circuit may be simultaneously varied. By arrangement of the transmitting stations signals can be transmitted in a desired direction.

1,514,661—M. W. Haub, filed November 6, 1922, issued November 11, 1924.

RADIO RECEIVING AND TRANSMITTING SYSTEM, comprising a circuit in which a condenser is connected in the antenna ground system and arranged with three sets of plates comprising a pair of dielectrically opposed plate members and a third set of plate members between the two sets of fixed plate members: The sets of fixed plate members have radio frequency energy impressed across them while the third set of plate members is connected with the lead from the grid of the electron tubes employed in the system.

1,514,699—E. C. Hanson, filed August 1, 1921, issued November 11, 1924.



NUMBER 1,514,699—Method and Apparatus for Radio Control for Torpedoes, and so on

METHOD AND APPARATUS FOR RADIO CONTROL FOR TORPEDOES, and so on, wherein a pair of loops are carried on aircraft and independently energized with audio frequency energy which is picked up by a tuned system having mechanical relays tuned to the audio notes transmitted. The signals of different notes are employed to actuate different controls aboard the torpedo.

1,514,732—P. J. Ruddy, filed June 21, 1921, issued November 11, 1924.

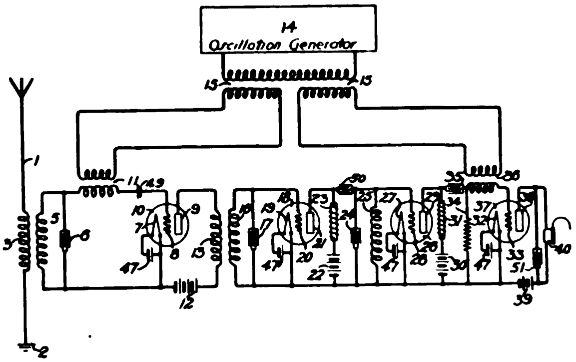
ELECTROMAGNETIC SIGNALING APPARATUS, in which the apparatus is self-contained and does not employ an antenna or ground connection. A device resembling a coherer is used at the receiver, consisting of a tube of non-conducting material, pole members at each end of the tube and groups of lead and magnetized steel balls between the pole members.

1,514,733—A. H. Sass, filed May 11, 1922, issued November 11, 1924. Assigned to Western Electric Company, Incorporated, New York.

CONDENSER, wherein both the movable and stationary plates are mounted on the same shaft, thereby decreasing the area occupied by the condenser. The shaft is insulated and arranged so that alternate plates are moved between the remaining plates which are immovable upon the shaft.

1,514,752—P. I. Wold, filed September 14, 1920, issued November 11, 1924. Assigned to Western Electric Company, Incorporated, New York.

Fig. 1



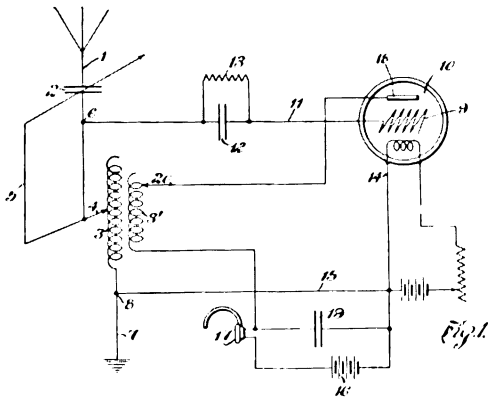
NUMBER 1,514,752—Method of and Means for Receiving Radio Signals

Method of and Means for Receiving Radio Signals, which comprises combining the incoming waves with locally generated oscillations of a frequency different from that of the incoming waves. The resultant composite waves are transferred selectively with the substantial elimination of the locally generated oscillations and then the resultant selectively transferred composite waves again combined with the locally generated oscillations reducing the frequency in such manner that it may be observed. The patent describes a single oscillator at the receiver which reacts a number of times upon the incoming signaling energy at different points in the amplification system and when the received energy has been modified in different conditions.

1,514,898—G. L. Geisey, filed July 18, 1923, issued November 11, 1924.

Thermionic Device, in which the electrodes are stamped from blanks with extended tongues or straps which enable the electrodes to be folded upon themselves and mechanically supported within the tube. The electrodes are punched with checkered apertures therein.

1,515,186—F. Conrad, filed June 3, 1920, issued November 11, 1924. Assigned to Westinghouse Electric and Manufacturing Company, Pittsburgh, Pennsylvania.



Number 1,515,186—Aperiodic Receiving System

Aperiodic Receiving System, employing inductive coupling with capacitance loading in the antenna circuit and in which automatic control of the antenna system is secured. The coupled

inductor and the condenser reactor are mechanically linked and varied simultaneously in the tuning of the receiver.

1,515,331—J. Bethenod, filed August 2, 1921, issued November 11, 1924.

RADIO TRANSMISSION SYSTEM, employing an antenna system which may be energized by a plurality of high frequency generators. All the generators may be operated simultaneously at the same frequency whereby the total generating power of the station is applied to the antenna, or these generators may be independently operated at different frequencies, in order to make possible multiplex transmission. The principle of the invention consists in that the antenna, which is a horizontal network extended in one direction, is divided into a plurality of sections each of which is supplied with current by a radio frequency generator, and provisions are made for modifying, if necessary, the effects of the electromagnetic or electrostatic induction between any two sections.

1,515,670—L. F. Fuller, filed September 25, 1919, issued Novembet 18, 1924. Assigned to Federal Telegraph Company, San Francisco, California.

RADIO TELEGRAPHY system, in which the ohmic resistance of the ground circuits is reduced with a view of raising the overall efficiency of a transmitting station. The specification points out that by reason of the large antenna current the I^2R losses in the usual ground system at a high power radio station have been so large as considerably to reduce the efficiency. By the present invention an ungrounded radiating circuit comprising a plurality of vertical loops in substantially the same plane is provided. The vertical sides of adjacent loops are near each other and current is supplied to each loop of such phase that the current in the adjacent vertical sides are opposed and substantially nullify each other. The resistance losses in the entire structure may be thus maintained at a desired low value.

1,515,900—C. J. Everett, filed May 5, 1922, issued November 18, 1924.

DETECTOR FOR USE IN RADIO CIRCUITS, in which the contact member comprises a spring positioned within a barrel and projecting out of the end thereof to engage the sensitive crystal surface.

1,515,990—R. D. Bangay, filed July 9, 1921, issued November 18, 1924. Assigned to Radio Corporation of America, Delaware.

RADIO TELEGRAPHY, in which the wave length of the transmitting system may be varied by movement of a variometer in the primary circuit. The primary inductance is in the form of a variometer while the antenna inductance is coupled with the variometer. By varying the angular position of the variometer coils in the primary circuit, the coupling is changed while maintaining the inductance constant.

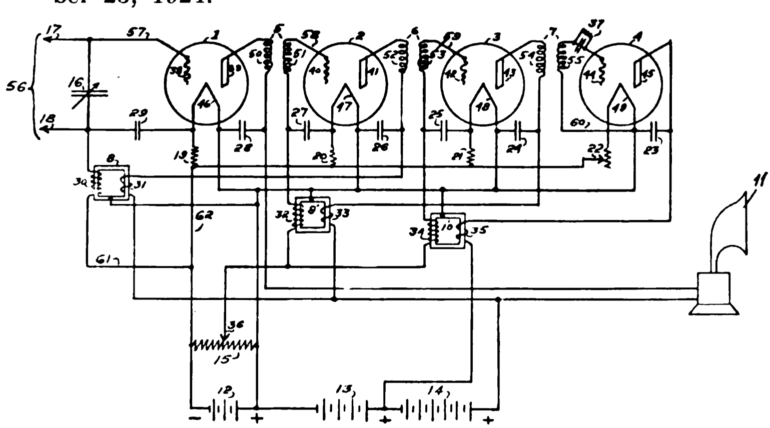
1,515,994—A. W. Bowman, filed April 4, 1923, issued November 18, 1924.

OSCILLATION DETECTOR of the crystal type, in which a crystal is disposed adjacent one end of a cartridge container while the surface is touched by a fine wire spiral spring projecting from the other end of the cartridge and controlled from a knob exterior of the barrel.

1,516,061—H. O. Rugh, filed November 16, 1922, issued November 18, 1924. Assigned to Rugh and Noble, Chicago, Illinois.

RADIO RECEIVING SYSTEM, which may be connected in the house lighting circuit similar to an incandescent lamp. A two-electrode tube is secured into the lighting socket and has its filament lighted from the source of current. A tuned circuit is provided across the input terminals while a responsive device is connected in circuit with the second electrode.

1,517,057—D. Grimes, filed September 19, 1922, issued November 25, 1924.

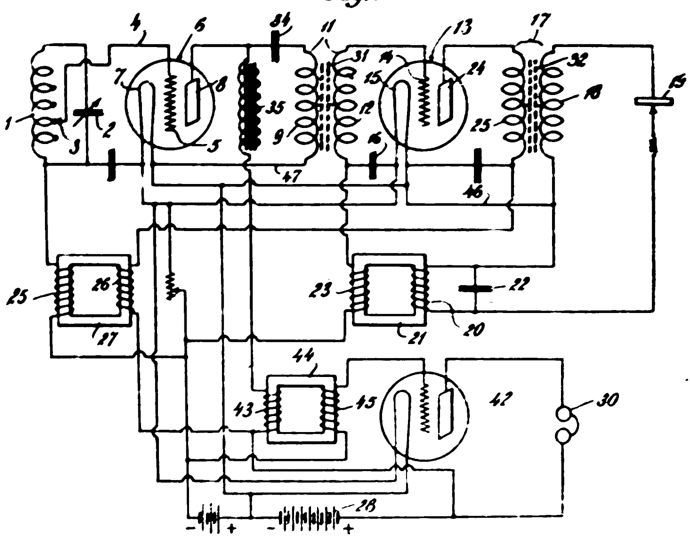


NUMBER 1,517,057—Vacuum Tube Amplifier

VACUUM TUBE AMPLIFIER, in which a plurality of tubes are provided so arranged that radio frequency current variations are repeatedly amplified by the electron tubes in a predetermined order with circuit connections whereby said tubes simultaneously operate repeatedly to amplify audio frequency current variations in the inverse order as compared with the predetermined order of radio frequency amplification.

1,517,058—D. Grimes, filed December 1, 1923, issued November 25, 1924. Assigned to Grimes Radio Engineering Company, Incorporated.

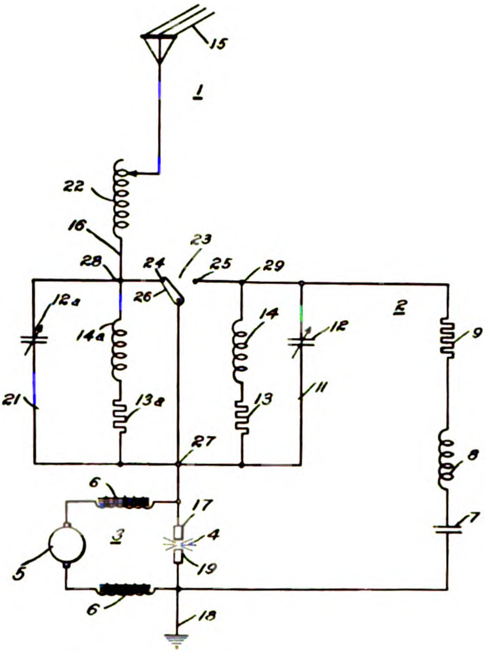
Fig. 6



NUMBER 1,517,058—Inverse Duplex Vacuum Tube Circuit

INVERSE DUPLEX VACUUM TUBE CIRCUIT, in which both radio and audio frequency amplification may be effected in tube circuits simultaneously. The radio frequency signaling currents are transmitted thru the amplification tubes in a certain order, while the audio frequency signaling currents are transmitted thru the same amplifier tubes successively in the inverse order with respect to the radio frequency amplification. Circuit connections are provided to prevent the effective transmission of audio frequency signaling currents from the output side of one amplifier to the input side of a successive amplifier in the order predetermined for radio frequency amplification.

1,517,277—O. B. Buchanan, filed August 18, 1921, issued December 2, 1924. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,517,277—Signaling System

SIGNALING SYSTEM for transmission by means of an arc on a single wave without interrupting the arc which consists in alternately inserting and removing parallel resonant reactance devices in the antenna circuit of the arc system for production of signals. The absorbing circuit is connected in parallel with the radiating circuit and each contains impedance elements having predetermined electrical time constants with switching means which render the sets of impedance elements effective or ineffective in accordance with the signals.

1,517,370—R. E. Marbury, filed November 26, 1920, issued December 2, 1924. Assigned to Westinghouse Electric and Manufacturing Company, Pittsburgh, Pennsylvania.

RADIO CONDENSER, in which the plates comprise metal foil coated with metal fusible below 200° C. The plates are alternately positioned between plates of solid dielectric. By this con-

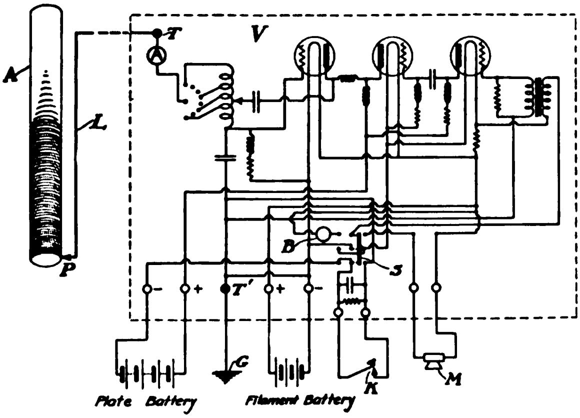
struction the condenser can be subjected to heat and great pressure so that the fusible metal will fill up all the irregularities in the dielectric sheets.

1,517,566—R. H. Marriott, filed November 1, 1921; issued December 2, 1924.

SPARK GAP unit for quenched gaps which has a spark chamber and a pressure relief chamber connecting therewith. The walls of the pressure relief chamber are expandible in such manner that on expansion of gas within the gap the unit will not develop leaks arising out of the tendency of the gas to escape from the unit.

1,517,569—J. O. Mauborgne and Guy Hill, filed June 3, 1921, issued December 2, 1924.

Fig. 2.

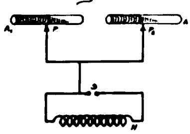


NUMBER 1,517,569—System of Radio Transmission

SYSTEM OF RADIO TRANSMISSION, in which a wave coil is employed in connection with a source of undamped high potential energy. A wave development is effected in the coil and the source of undamped oscillations modulated for the production of signals.

1,517,568—J. O. Mauborgne and Guy Hill, filed June 16, 1920' issued December 2, 1924.

Fig. 7.



NUMBER 1,517,568 —System of Radio Transmission

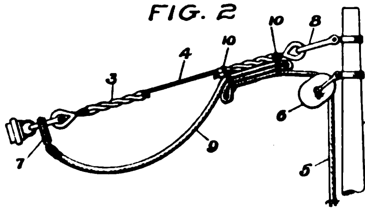
System of Radio Transmission which includes a power supply and two or more wave coils. Corresponding points are selected on the coils and a suitable high frequency potential from the power supply impressed between the points for causing wave developments on the wave coils. The power supply is modulated for the purpose of signaling.

1,517,570—J. O. Mauborgne and Guy Hill, filed February 17, 1921, issued December 2, 1924.

System of Radio Communication utilizing a wave coil as an antenna. An elongation of variable length is provided on one end of the wave coil and the capacity of the coil and the wave distribution thereon are thereby varied.

1,517,602—A. M. Trogner, filed July 30, 1920, issued December 2, 1924.

FIG. 2



Number 1,517,602—Antenna Safety Link

Antenna Safety Link having a weakened portion therein, which may be ruptured when the antenna wires are subjected to undue strain without injury to the antenna conductors. The object of the invention is to prevent the carrying away of the antenna as the result of an abnormal "whipping" of the masts on shipboard, such as happens in cases of collision, torpedoing, grounding, and the like.

1,517,654—H. J. Round and A. McLellan, filed March 30, 1921, issued December 2, 1924. Assigned to Radio Corporation of America, New York.

Radio Signaling System, in which the amplitude of the energy emanating from a transmitting station may be maintained constant. A tube transmitter is shown having an oscillation circuit and connections for impressing the oscillations generated on an antenna system. A relay is energized by energy delivered to the antenna system, so that upon tendency of the energy to vary, the relay operates to prevent the generated oscillations from falling below a predetermined value.

1,517,816—E. F. W. Alexanderson, filed June 24, 1921, issued December 2, 1924. Assigned to General Electric Company, New York.

RADIO TRANSMITTING SYSTEM for a multiple tuned antenna divided into different portions with circuits for supplying the signaling energy to the different portions and varying the phase of the current so supplied to obtain maximum radiation of the signaling energy.

1,518,439—A. Meissner, filed September 3, 1921, issued December 9, 1924. Assigned to Gesellschaft für drahtlose Telegraphie, m.b.H., Halleschès, of Berlin, Germany.

ARC TRANSMITTER FOR RADIO TELEGRAPHY, in which the arc generator is connected with the antenna system thru a circuit arranged to increase by an even or odd multiple of the frequency of the oscillations, the frequency of the radiated signaling energy. The patent discloses a magnetic frequency changer interposed between the arc generator and the radiating system.

1,518,564—T. S. Cole, filed October 27, 1922, issued December 9' 1924.

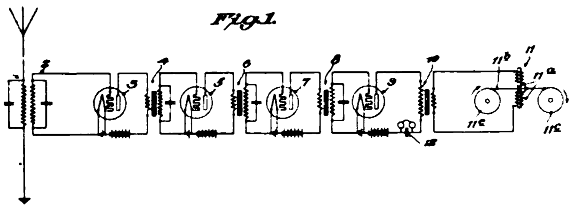
RADIO TELEPHONE AND TELEGRAPH APPARATUS, consisting of a battery system for connection in the plate circuit of an electron tube with a switching arrangement associated with the battery system, whereby in periods of non-use of the radio receiving circuit the "B" battery may be connected in shunt with the "A" battery for enabling the "B" battery to be charged from the "A" battery. The "B" battery is constructed in the form of a secondary battery having its sections so arranged in parallel thru the switching arrangement that the filament battery may be utilized to charge the plate battery.

1,518,633—R. E. H. Carpenter filed May 20, 1924, issued December 9, 1924.

RADIO SIGNALING SYSTEM AND APPARATUS THEREFOR for the reception of signaling energy in which the receiver may be accommodated to various antenna systems and to suit the characteristics of such antenna systems to which the apparatus may be connected. A coupling system is provided between the receiver and the antenna circuit which consists of a resistance connected in series with a tuning condenser and in shunt with the antenna circuit. The resistance and condenser are varied in order to tune

the receiving system for operation and connection with antennas of different characteristics.

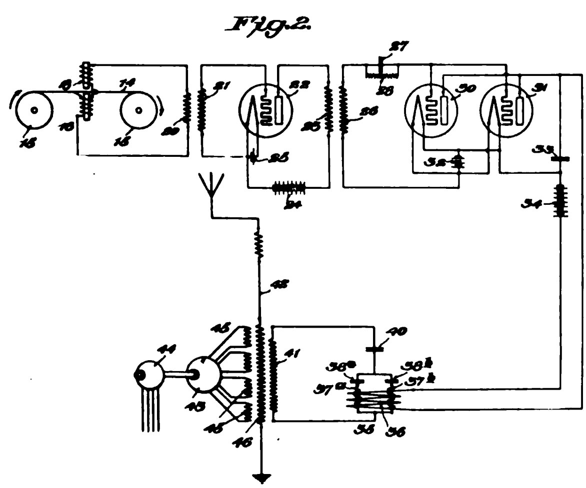
1,518,655—W. L. Carlson and E. C. Hanson, filed January 14, 1920, issued December 9 1924.



Number 1,518,655—Radio Telegraph System

Radio Telegraph System employing the telegraphone as a recorder of incoming signaling energy. An electron tube amplification circuit is provided, the output circuit of which contains a device for audibly reproducing the signals and also the recording heads of the telegraphone whereby a magnetic record of the incoming signals is simultaneously produced.

1,518,656—E. C. Hanson, filed March 11, 1924, issued December 9, 1924.



Number 1,518,656—Radio Telegraph System

Radio Telegraph System employing a telegraphone upon the wire of which the signals are initially impressed and then the

telegraphone operated at high speed to control the radiation of signaling energy from a source of sustained oscillations. The invention is directed to automatic high speed signaling from high power stations.

1,518,682—W. R. G. Baker, filed November 6, 1923, issued November 9, 1924. Assigned to General Electric Company of New York.

SIGNALING SYSTEM, in which an electron tube is employed as a generator of oscillations and is maintained in the condition of oscillation continuously while signals are produced by switching the oscillator alternately to a storage circuit and the radiating circuit. The system is intended principally for tubes of high power which will not permit the keying of the grid circuit.

1,519,398—H. F. Elliott, filed December 14, 1921, issued December 16, 1924. Assigned to Federal Telegraph Company, San Francisco, California.

ARC CONVERTER, including an arc chamber and a mechanical arrangement for exhausting and blowing out the arc chamber with a valve arranged to connect either the exhausting pump or the blower with the arc chamber. The arc is operated in an atmosphere of hydrogen or other gas having similar electromechanical properties. The apparatus disclosed in this patent is in association with the arc converter for controlling the condition thereof.

1,519,412—S. R. Mullard, filed October 2, 1922, issued December 16, 1924. Assigned to The Mullard Radio Valve Company, Limited, England.

SUSPENSION OF INCANDESCENT FILAMENTS for electron tubes for ensuring that the filament will be maintained under tension at all times. The filament is supported in spring suspension within the helix formed by the grid and inside of the vertical plate electrode.

1,519,615—R. A. Heising, filed June 5, 1920, issued December 16, 1924. Assigned to Western Electric Company, Incorporated, New York.

SIGNALING SYSTEM having a modulating system of the type wherein the unmodulated component of the modulated carrier frequency current is wholly or partly suppressed. An oscillator is provided for generating modulated current. A plurality of

paths are arranged for the passage of this current and one of the paths includes a stiffly resonant circuit of zero reactance and resistance for the unmodulated component of the modulated current. Another of the paths includes a loosely resonant circuit tuned to the same frequency. The undesired side frequencies are passed into the branch paths while the main frequency is transmitted.

1,519,899—R. C. Benner, filed February 28, 1922, issued December 1, 1924. Assigned to National Carbon Company, Incorporated, New York.

APPARATUS FOR RADIO COMMUNICATION, comprising a construction of "B" battery. The "B" battery comprises a plurality of cells each having a substantially flat zinc anode which is disposed parallel and in close relationship with another zinc anode of the succeeding cell. The plates are substantially overlapped. The battery, therefore, has a large concentrated capacitance and according to the specification the electron tube circuit becomes more sensitive to weak radio frequency currents.

1,510,027—A. A. Kent, original filed September 23, 1916, issued December 23, 1924. Assigned to Atwater Kent Manufacturing Company, Philadelphia, Pennsylvania.

CONDENSER AND HOLDER THEREFOR, in which the condenser is gripped between a pair of pressure plates which form the holder for the condenser adjacent a pair of binding posts comprising the terminals of the condenser.

1,520,329—C. S. Cherpeck, filed August 26, 1922, issued December 23, 1924.

VARIABLE CONDENSER of the book type in which plates which are normally paralleled may have their capacity varied by angularly varying the pistons between the plates by means of a screw which may be advanced to introduce a cam movement arranged to vary the angularity of the plates.

1,520,461—H. A. Bremer, filed September 21, 1923, issued December 23, 1924.

ELECTRIC CONDENSER, in which the plates are arranged in pairs and are dished in such manner that movable plates may be interleaved between the stationary plates, the offset portions of the movable plates moving between the offset portions of the stationary plates. In this way a smaller number of spacing mem-

bers is required in the support of both the movable plates and the stationary plates.

1,520,580—Norman Lea and John Ree, filed September 26, 1919, issued December 23, 1924.

Electromagnetic Wave Signaling System, consisting of an electron tube transmitter wherein the oscillation circuit has a permanent conductive connection between the control element of the valve and the negative lead of the high tension supply. The signals are produced by breaking the lead between the filament and the negative lead to the control electrode.

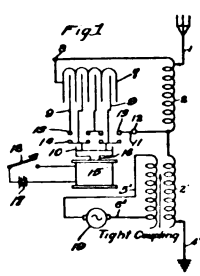
1,520,640—G. L. Geisey, filed November 20, 1922, issued December 23, 1924.

Thermionic Device or electron tube in which the grid is formed by a checkered plate with spaced openings while the plate is formed from flat material having raised portions in checkered formation corresponding to the checkered spaces in the grid. The tube is constructed in such manner that the electrodes are spaced very close together.

1,520,835—A. Meissner, filed September, 1921, issued December 30, 1924. Assigned to Gesellschaft für drahtlose Telegraphie, m.b.H., of Berlin, Germany.

Method of Receiving Electrical Oscillations, which includes heterodyning the received energy by a local source of current which has a frequency differing by an audio frequency from a multiple of the received frequency. The object of this invention is to provide a selective system of reception by choosing a particular frequency for the local source.

1,521,018—H. L. Godfrey, filed June 3, 1920, issued December 30, 1924. Assigned to Westinghouse Electric and Manufacturing Company, Pennsylvania.



Number 1,521,018—
Signaling System

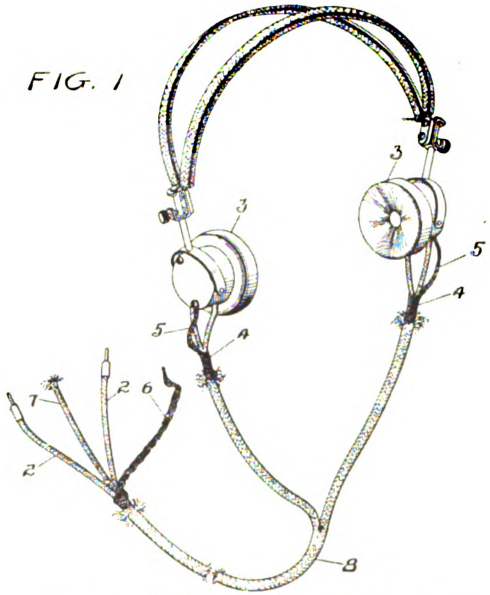
SIGNALING SYSTEM, in which the transmission of the signaling energy is controlled by opening and closing a frequency trap circuit. An inductor at the transmitter has a condenser connected in parallel therewith. The condenser is constructed of a plurality of units which may be simultaneously shunted in forming the signals.

1 521,205—W. S. Stephenson and G. W. Walton, filed March 24, 1924, issued December 30, 1924.

SYNCHRONIZING ROTATING BODIES at radio transmitting and receiving stations intended for operation of apparatus in the transmission and reception of photographs. Synchronizing signals are transmitted and received in such manner that they will control the operation of rotary apparatus at the receiver for maintaining the receiving apparatus in step with the transmitting apparatus for purposes of synchronizing the photograph process.

1,521,275—G. W. Carpenter and W. L. Carlson, filed January 29, 1921, issued December 30, 1924.

FIG. 1

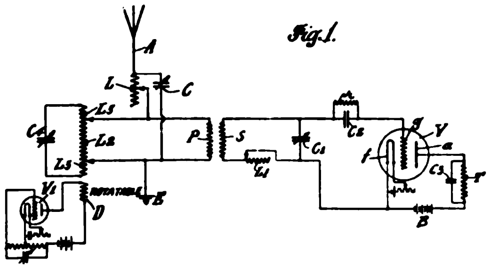


NUMBER 1,521,275—Telephone Headset

TELEPHONE HEADSET where the cords of the telephone headset are electrostatically shielded by a flexible conductive plate, which consists of woven tinsel conductors. The shield extends

over the telephone conductors and is grounded on the caps of the telephone receivers. The shielded headset is described for use with sensitive multi-stage electron tube amplifiers in long distance reception.

1 521,380—D. G. McCaa, filed November 17, 1922, issued December 30, 1924. Assigned to The Electric Apparatus Company, Parkersburg, Pennsylvania.



NUMBER 1,521,380—Receiving System

RECEIVING SYSTEM for radio signals in which a circuit arrangement is provided for discriminating against static atmospherics, strays and other disturbances. The received energy representing both the desired signal and the disturbing effect is divided into two paths, including reactive devices, one of which is employed for effecting the translation of the desired signals and with another of which is associated a local source of oscillations in such manner as to cause the effect of said reactance to fluctuate within wide limits resulting in the fluctuation in amplitude of the signal representing energy in the first reactance and at certain instants to be reinforced by energy from the local source to the substantial exclusion or great reduction of the effects of the simultaneously existing disturbing energy.

NUMBER 66,049—
Advertising Support for Telephone Head Sets

66,049—L. J. Urich, filed September 3, 1924, issued November 18, 1924. Assigned to C. Brandes, Incorporated, of New York, N. Y.

ADVERTISING SUPPORT FOR TELEPHONE HEAD SETS, in which a window display statuette is arranged with indentations formed on the head of the statuette to receive and support a telephone headset for display purposes. The statuette has the appearance of a radio listener-in wearing the headset which is on display.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 13 APRIL, 1925 Number 2

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CORRECTION

On page 805 of THE PROCEEDINGS for December, 1924 (Volume 12, Number 6), in footnote 3, line 4, change "Major Le Fry" to "Major Lefroy."

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The PROCEEDINGS of the Institute are published every two months and contain the papers and the discussions thereon as presented at the meetings in New York, Washington, Boston, Seattle, San Francisco, Philadelpha, or Chicago.

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SOME TRANS-PACIFIC RADIO FIELD INTENSITY MEASUREMENTS*

By

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(Conducted jointly by the Bureau of Standards and the American Section of the International Union for Scientific Radiotelegraphy.)

Radio field intensity measurements for frequencies varying roughly between 1,000 and 15 kc. (300 m. and 20,000 m.), and for distances up to 6,500 km. by daylight and over salt water, have been made by a number of independent observers and the results with some exceptions[1] are in fair agreement. For frequencies from 1,000 at least down to 60 kc. (300 m. to 5,000 m.), the observed results agree within the limits of experimental certainty with the values calculated from the Austin-Cohen formula, up to the greatest distances attempted (5,500 km.).[2] The lower frequencies ordinarily used for long-distance communication, say from 15 to 30 kc. (10,000 to 20,000 m.), give observed values somewhat larger than those calculated. At a distance of 6,000 km. this ratio of observed calculated values amounts, on an average, to about two to one.[3] Only a limited number of observations have been taken at distances much greater than this, and these have generally indicated a considerable increase in the observed to calculated ratio.[4]

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*Received by the Editor, January 24, 1925. Published by permission of the Director of the Bureau of Standards of the United States Department of Commerce.

[1] G. Vallauri, Proceedings of The Institute of Radio Engineers volume 8, page 286, 1920.
M. Guierre, "Radio Review," volume 2, page 618; 1921.
M Baumler, "Elek. Nach. Tech.," volume 1, page 50; 1924.

[2] L. W. Austin, "Bureau of Standards Scientific Papers, number 159, 1911, and number 286, 1914.
R. Bown, C. R. Englund and H. T. Friis, Proceedings of The Institute of Radio Engineers, volume 11, page 115, 1923.

[3] G. W. Pickard, Proceedings of The Institute of Radio Engineers, volume 10, page 161, 1920.
J. L. Eckersley, "Jour. I. E. E." (London), volume 58, page 677, 1921.
L. W. Austin, Proceedings of The Institute of Radio Engineers, volume 11, page 459, 1923 and volume 12, page 389, 1924.

[4] M. Guierre, "Radio Review," volume 2, page 618, 1921.

In order to get more reliable data at these greater distances, observations were taken in August and September, 1924, at San Diego, California, on the signals from Cavite, Philippine Islands, and from Malabar, Java, the distance from Cavite to San Diego is 11,000 km. with a difference in time of eight hours. This gives about two hours for observations in September without approaching the time of sunrise or sunset too closely. The distance from Malabar is 14,700 km., with a time difference of nine hours. These are about the greatest distances that can be obtained for all daylight and approximately all water signal path with the present high-power stations of the world, except perhaps between Japan and Chili. This last would, however, give about ten hours difference in time and bring the sending and receiving stations rather close to their respective sunrise and sunset times.

The receiving measurements were made in the United States Naval receiving station, Point Loma, San Diego, under receiving conditions which, while not ideal, are believed to have given errors not greater than twenty per cent. On account of the weakness of the signals, in comparison with the atmospheric disturbances, and to keep out strong interference from eastern stations, it was generally found necessary to make the measurements on Cavite and Malabar with uni-directional reception, using the general type of circuit described in the work on the direction of atmospheric disturbances in 1920.[5]

The arrangement of circuits is shown diagrammatically in Figure 1. Here *A* is a single-wire antenna approximately 30 m. long and 20 m. high, *B* a square coil antenna of 48 turns and 2.44 m. on a side. This was mounted so as to be capable of rotation, with its lower side about 3 m. from the ground. *C* is an intermediate circuit to reduce interference, *D* is the detector circuit, and *E* represents three stages of radio-frequency transformer coupled amplification adjusted to prevent regeneration. *F* is a heterodyne generator for producing the local oscillations for beat reception. *H* is a telephone comparator,[6] consisting of a General Radio tuning fork oscillator which generated a thousand cycle current which was measured on a thermo-galvanometer and then passed through a voltage divider and resistances so that a known 1,000-cycle emf. could be impressed on the telephones. In making the measurements, the heterodyne was adjusted so as to give a beat frequency with the signal equal to

[5] L. W. Austin, "Jour. Franklin Institute," page 619, 1921.

[6] Austin and Judson, Proceedings of The Institute of Radio Engineers, volume 12, page 521, 1924.

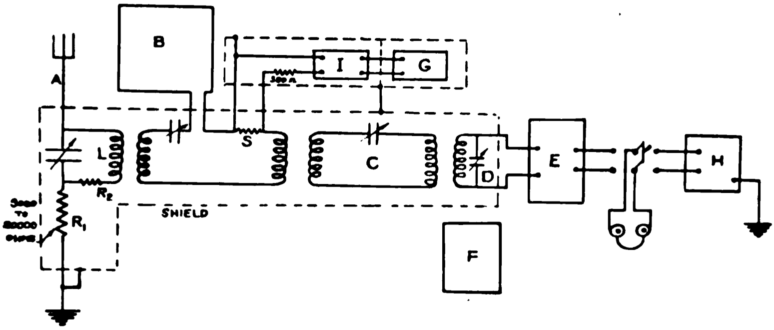

FIGURE 1

the frequency on the telephone comparator. The telephones could than be switched rapidly from the receiver to the telephone comparator, and the voltage divider adjusted until the telephone note was of the same intensity in each. The couplings between *B*, *C*, and *D* were all loose and remained fixed during the whole course of the experiments. The heterodyne was coupled to the detector so as to give the loudest signal, and the coupling was separately adjusted for the different wave lengths received. The inductance *L* was connected to the antenna thru a reversing switch so that the elevated antenna coil combination could be made to receive from either direction in the plane of the coil, while reception from the opposite direction was practically zero.

The receiving set was calibrated by introducing a known emf. in the coil antenna from the radio-frequency generator *G*. This consisted of a tuned plate electron tube generator proper, the output current of which, after being measured with a thermoelement and galvanometer passed thru an attenuation box (artificial line) *I*, kindly loaned by the Western Electric Company. From this it passed to a 1-ohm resistance *S* inserted in the loop. Ordinarily the current from *G* was adjusted to 1.6 milliamperes, which was reduced in the attenuation box *I* to 1/500th of its value, thus giving 3.2 microamperes in the resistance *S*. To prevent capacity coupling, the coil antenna was grounded at one terminal of *S*. The generator with its dry cell batteries and the attenuation box were enclosed in grounded copper boxes, as was the whole receiving set with the exception of the coil antenna.

The calibrating generator was furnished with a fixed condenser giving a frequency of 23.06 kc. (13,000 m.). It had been intended to replace this in San Diego by a variable condenser

so that calibration could be made at all the frequencies of the stations being measured. Unfortunately no variable condenser was found available which was small enough to be used in the copper box, so that it was necessary to do the calibrating at the single wave length and make corrections for the effect of change of frequency. As the calibration was made on the coil antenna with the elevated antenna disconnected, a correction for the presence of the elevated antenna in the reception of the signals was also necessary. This correction was determined by measurements made on the stronger stations with and without the elevated antenna. Reversal experiments with the coil antenna showed that the effect of its capacity to earth was negligible.

Table I gives the date for calculation of the field intensities of the sending stations. The Malabar antenna[7] is suspended by steel cables in a mountain ravine about 1.5 km. wide at the top and with an average depth of about 550 m. According to information kindly furnished by Mr. Schotel of the Dutch Colonial Office, the antenna current is approximately 500 amperes at the frequency measured in San Diego. Estimates of the radiation height, from measurements made at moderate distances, vary between 320[8] and 480 m. These varying results are probably due to the mountainous character of the surrounding country. In the calculations the lower value, 320 m. has been used. Even with this value the observed to calculated ratio of Malabar at San Diego is considerably less than that of Cavite, notwithstanding its greater distance. This may be due to the mountainous surroundings.

TABLE I

Transmitting Station Data

	Pearl Harbor NPM	Tuckerton WGG	Cavite NPO	Malabar PKX
Frequency, kc.	24.80	18.86	19.34	18.98
Wave length, m.	12,090	15,900	15,500	15,800
Antenna current, Amp.	170	470	180	500
Radiation height, m.	120	67.5	120	320
Distance, km.	4,200	3,800	11,800	14,700

[7] C. J. DeGroot, Proceedings of The Institute of Radio Engineers, volume 12, page 693, 1924.

[8] Other engineers give even lower values.

Table II shows the field strength in microvolts per meter for 1 ohm on the voltage divider of the telephone comparator at the various frequencies observed, with the amplifier adjusted so that 3.2 microvolts in the coil B gave 50 on the telephone comparator.

TABLE II

TELEPHONE COMPARATOR FACTOR (A) FOR VARIOUS FREQUENCIES

f (kc.)	λ (m.)	A
24.80	12,090	0.402
19.34	15,500	0.504
18.98	15,800	0.524
18.86	15,900	0.545

Comparator reading $\times$ A = Electric field intensity in microvolts per meter.

On account of an accident to the calibrating apparatus and delays in getting replacements, the first part of the work had to be confined to comparisons of Cavite (NPO) and Malabar (PKX), with the stronger stations, Pearl Harbor (NPM) and Tuckerton, New Jersey (WGG). The Cavite and Malabar average readings were later reduced to microvolts per meter, assuming that the average strengths of Pearl Harbor and Tuckerton were the same as their averages during the second period when the calibrations were being made.

Table III gives the comparator readings and the ratios of the various stations for the first period of the observations, while Table IV shows the results during the second period when the field intensities were measured directly.

The observations contained in the tables were all taken between two and four o'clock in the afternoon, Pacific time. A few observations taken in the morning when the signal path was partly in daylight and partly in darkness, indicated a somewhat greater strength than the afternoon observations; while observations taken by the station operators at the time of darkness along the whole signal path, were reported to be many times stronger than the daylight observations, approaching at times the strength of Pearl Harbor (NPM).

The final values of the field strengths of Cavite and Malabar as derived from comparison with the signals of Tuckerton and

Pearl Harbor, and by direct measurement, are shown in Table V. Below the average observed values are given the values calculated from the Austin-Cohen formula,

TABLE III

Comparisons of Cavite (NPO) and Malabar (PKX) with Tuckerton (WGG) and Pearl Harbor (NPM)

1924	Telephone Comparator Readings				Telephone Comparator Ratios				Disturbances $\mu v./m.$
	NPH	PGG	NPO	PXX	$\frac{NPM}{NPO}$	$\frac{WGG}{NPO}$	$\frac{NPM}{PKX}$	$\frac{WGG}{PKX}$	
Aug. 28	85	....	7.5	....	11.3	....	....	....	14
30	110	....	6.5	13.0	16.9	....	7.4	....	23
Sept. 1	55	....	3.5	4.0	15.7	....	13.7	....	....
2	....	45	....	5.0	....	....	....	9.0	12
3	....	40	....	3.0	....	....	....	13.3	....
4	....	100	....	17.0	....	....	....	5.9	38
5	....	100	3.0	....	....	33.4	....	....	....
6	100	100	6.3	....	15.8	15.8	....	....	....
8	30	80	4.3	....	11.6	18.6	....	....	50
9	55	80	4.3	7.7	12.8	18.6	7.1	10.4	25
10	45	80	5.0	6.0	9.0	16.0	7.5	13.3	30
11	50	78	4.3	....	11.6	18.1	....	....	....
12	40	70	3.0	10.0	13.3	23.4	4.0	7.0	5
13	50	80	6.3	....	7.9	12.7	....	....	20
15	48	80	....	17.0	....	....	2.8	4.7	25
Average					12.59	19.60	7.05	9.08	

TABLE IV

Direct Measurements Using the Calibrating Generator

Sept. 1924	Pearl Harbor (NPM)		Cavite (NPO)		Tuckerton (WGG)		Malabar (PKX)		Atmospheric disturbances $\mu v/m.$
	Telephone Comp.	$E \frac{\mu v}{m}$	Telephone Comp.	$E \frac{\mu v.}{m.}$	Telephone Comp.	$E \frac{\mu v.}{m,}$	Telephone Comp.	$E \frac{\mu v.}{m.}$	
17	40	16.1	4.0	2.0	....	....	7.1	3.7	14
			5.0	2.5					
			4.0	2.0					
18	45	18.1	6.0	3.0	90	49.0	4.8	2.5	
	60	24.2	3.0	1.5					5–10
19	60	24.2	....	....	100	54.4	7.5	3.9	5–10
20	30	20.1	4.0	2.0	50	27.2	....	....	15
	30	12.0	6.0	3.0	70	38.1			
	40	16.1			60	32.6			
					70	38.1			
22	60	24.2	4.0	2.0	70	38.1	13.0	6.8	5–10
					90	49.0			
Average	48.3	19.4	4.5	2.27	75	40.9	8.1	4.2	

$$E = \frac{377\,I\,h}{\lambda d}\sqrt{\frac{\theta}{\sin\theta}}\,\varepsilon^{-u} \qquad (1)$$

where $u = 0.0015d/\sqrt{\lambda}$ and θ is the angle between the stations from the center of the earth.

The observed value of field intensity in the case of Cavite in Table V is seen to be approximately three times that calculated from equation (1), while the observed strength of Malabar is about twice that calculated from equation (1). These ratios of observed to calculated values are much less than those given in M. Guierre's paper.[8]

TABLE V

AVERAGES OF FIELD INTENSITY

	Cavite $E\frac{\mu v.}{m.}$	Malabar $E\frac{\mu v.}{m.}$
From direct measurements..........	2.27	4.22
From comparison with NPM........	1.93	3.55
From comparison with WGG........	1.93	4.26
Observed averages...............	2.04	4.02
Calculated from equation (1).....	0.69	1.83

SUMMARY: The paper describes measurements on the daylight radio field intensity produced in San Diego, California, by the arc stations at Cavite, Philippine Islands, 11,800 km., and Malabar, Java, 14,700 km. distant. These distances are nearly twice as great as any previously studied, except for a few scattered measurements. The average observed intensities were, from Cavite 2.04 $\mu v./m.$, and from Malabar 4.02 $\mu v./m.$, while those calculated from the Austin-Cohen formula are respectively 0.69 $\mu v./m.$, and 1.83 $\mu v./m.$ These ratios of observed to calculated values indicate an increase in the divergence from the formula with increasing distance, but not so great as was indicated by earlier scattered observations.

THE MAGNETRON AMPLIFIER AND POWER OSCILLATOR*

By

Frank R. Elder

Introduction

It is well known that there are two general methods of affecting the motion of charged particles, namely by an electrostatic field or by a magnetic field. Therefore in electron tubes these principles may be employed to control the flow of electrons. As an example of the first type we have the ordinary three-electrode tube or radiotron. The second type is exemplified by the magnetron. There are two kinds of magnetrons. In the first the magnetic field coil which acts as the control member is external to the tube, while in the second the magnetic field of the current that heats the filament is the control agent.

The radiotron has come into quite general use and its capabilities have become well known. The magnetron is still in its infancy, however, and comparatively little has been written about it. The magnetron has been described and some of its uses briefly mentioned in a paper by A. W. Hull.[1] The axially-controlled magnetron has been described by the same author.[2] It is the purpose of the present paper to set forth in detail some of the data which have accumulated in this laboratory on two applications of the magnetron, first as an amplifier and second as a power oscillator.

Design of Tubes

The magnetron is a vacuum tube with two elements, a filament and an anode, like a simple kenotron, its only distinguishing feature being the symmetrical arrangement of its electrodes. As generally made, the filament was a single strand of wire in the axis of the anode (Figure 1). A spiral spring in series with the filament served to keep the filament straight when heated. The anode, which was usually of molybdenum, was spaced from the glass wall of the container by spirals of fine molybdenum or tungs-

* Received by the Editor, September 20, 1924.

ten wire attached near the ends. In order to prevent eddy current losses in the anode as much as possible, the anode was split longitudinally. The anode lead was sometimes brought out thru the side as shown in Figure 1, or at the end as shown in Figure 2. In the second case the anode lead had to be protected by a glass tube (or a quartz tube if high voltage was to be applied to the anode), otherwise the symmetry of the device was spoiled and the magnetic control made very poor. It was also observed,

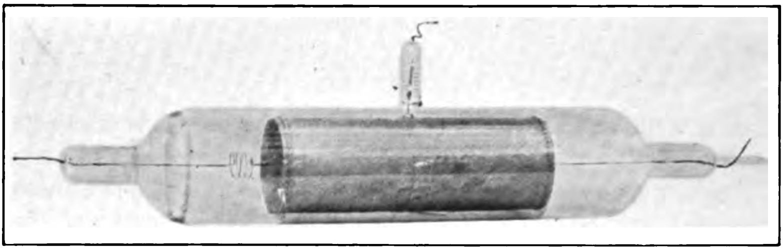

FIGURE 1—Magnetron[4]

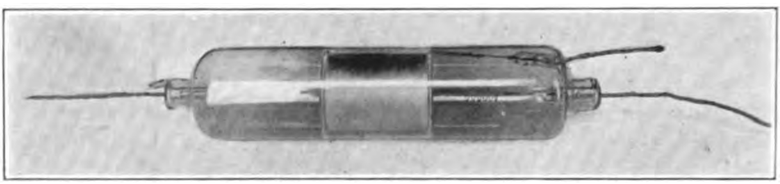

FIGURE 2—Magnetron

when high voltage was applied to the tube of the form shown in Figure 1, that tiny bright arcs frequently appeared, when the magnetic field was applied, at the points of contact of the supporting spirals with the glass wall. This led to the construction shown in Figure 3 where the anode was supported by conical spiral springs terminating in quartz insulators.

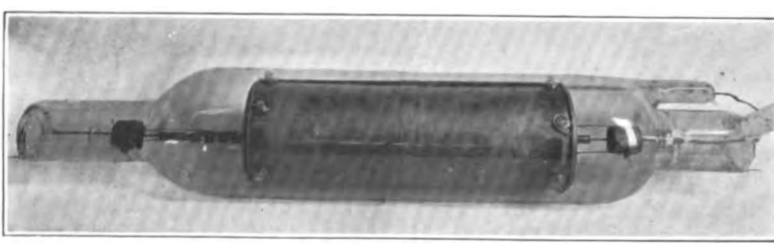

FIGURE 3—Magnetron

Many sizes of tubes have been made and studied; a partial list follows:

I	II	III	IV
¾×1½″	1″×4″	1″×12″	1¼″×1¾″
1.9×3.31	2.54×10.16	2.54×30.5	3.17×4.45 cm.
V	VI	VII	VIII
2″×2″	2″×4″	2″×6″	4″×12″
5.08×5.08	5.08×10.16	5.08×15.25	10.16×30.5 cm

The first dimension given is the anode diameter, the second dimension the anode length. The sizes marked IV and V were used chiefly as amplifiers while all sizes were studied as oscillators. With the first a few watts output at about 180 meters was obtained; with the last several kilowatts at 10,000 meters.

THEORY

In the absence of a magnetic field the current which can flow between a cathode and a concentric cylindrical anode tube is given by the relation [3] [4].

$$i = 14.65 \times 10^{-6} \frac{l}{r\beta^2} V^{\frac{3}{2}} \qquad \text{(I)}$$

where i is the current in amperes, l the length, and r the radius of the anode, V the voltage of the anode, and β^2 a factor depending upon the ratio of the radii of anode and cathode. For practical tubes β^2 is approximately unity.

If now a uniform magnetic field is applied parallel to the axis of the tube, this current is not affected until a certain critical value of field H is reached when the current falls abruptly to zero. The relation between voltage and field has been shown to be [5

$$V = 8 \frac{e}{m} r^2 H^2$$

where V is the voltage of the anode, r the radius of the anode, and H the applied field, e the charge, and m the mass of an electron in e. m. units. In practical units (volts, cm., gausses), this equation is $V = 0.0221\, r^2 H^2$ (11)
This may also be written in the form

$$H = \frac{6.73}{r} V^{\frac{1}{2}} \qquad \text{(III)}$$

That this is the correct relationship has been abundantly verified by experiment. A series of characteristics taken on a 4″×12″ tube is shown in Figure 4. There may be a small residual current with strong field due to electrons reaching the anode after colliding with gas molecules. The steep portion of the curve is not quite vertical but becomes more nearly so as the voltage

is increased. For moderate voltages a 10-percent change in magnetic field suffices to change the current from the full space charge value of equation (1) to practically zero.

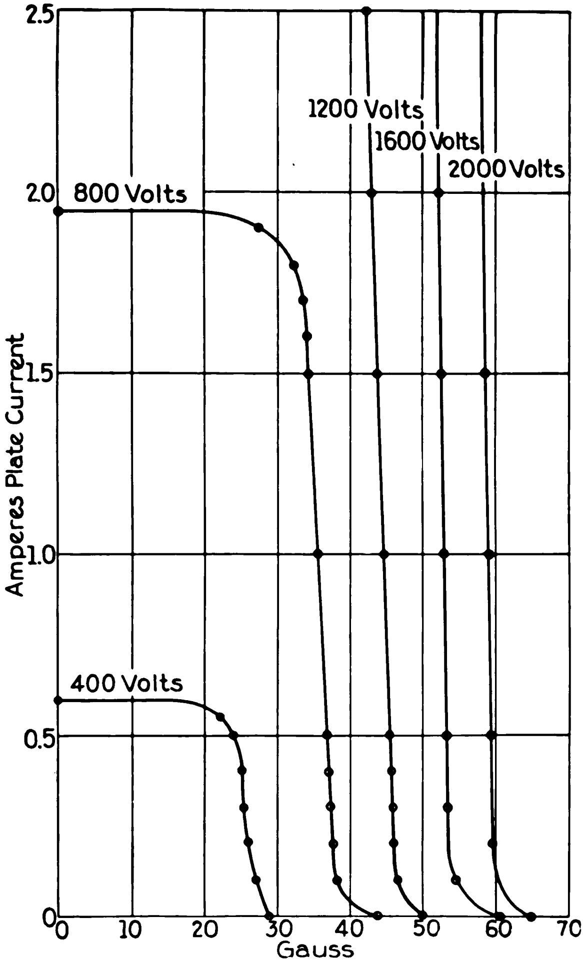

FIGURE 4—Characteristic Curves 4"×12" Magnetron

PART I

THE MAGNETRON AMPLIFIER

Most of the amplification tests were carried out at a wave length of 8,000 meters, and the discussion will be limited to the results obtained at this wave length.

THE CIRCUIT

The circuit used in these tests is shown in Figure 5. The field

coils serve to "polarize" or "bias" the plate current to some point on the steep part of the characteristic curve (Figure 4), usually rather well down on the curve in order to limit heating of the anode. At L_1 and L_2 are shown the control coils by which the signal is impressed on the tubes, C and C_1 are tuning condensers which are variable, and C_2 is a stopping condenser to prevent short circuiting of the plate voltage in case the insulation of C_1 fails. Between the plate voltage supply and the plate of the tube is a choke coil to prevent the radio frequency going to ground thru the plate battery or generator.

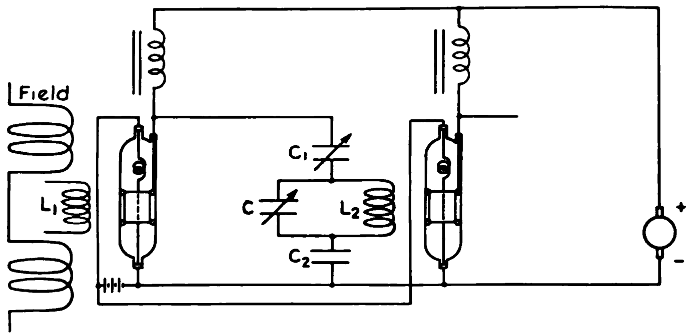

FIGURE 5—Circuit for Magnetron Amplifier

In order to obtain the best results the impedance of the output circuit must be made to fit the output impedance of the tube. Let us consider briefly the result of impressing a voltage $E \sin \omega t$ across the combination C, C_1, L as represented in Figure 6. A

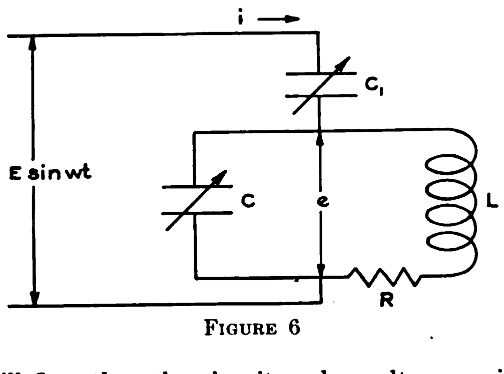

FIGURE 6

current i will flow thru the circuit and a voltage e will be set up across C and L. As is usually the case with radio circuits, the resistance of the coil L can be neglected in comparison with ωL, and thus in the complex notation,

$$\frac{i}{e}=\frac{1}{\omega L}\cos\theta+j\,\omega\left(C_1+\frac{1}{\omega L}\right) \qquad \text{(IV)}$$

where $$\cos\theta=\frac{R}{\sqrt{\omega^2 L^2+R^2}}$$

Inverting equation (4) we have

$$\frac{e}{i}=\frac{\frac{1}{\omega L}\cos\theta-j\left(\omega C_1-\frac{1}{\omega L}\right)}{\left(\frac{1}{\omega L}\cos\theta\right)^2+\left(\omega C_1-\frac{1}{\omega L}\right)^2} \qquad \text{(V)}$$

which is of the form

$$\frac{e}{i}=X+j\,Y \qquad \text{(VI)}$$

where X represents the total impedance of the circuit and Y represents the inductive reactance of the combination C and L, which is to be tuned out by C_1, when C is a capacity less than that required to tune L to resonance at the frequency ω. It is thus seen that the circuit will act as a pure resistance of any desired value X. Reduction of equation (V) leads to

$$C=\frac{\frac{1}{\omega L}-\sqrt{\frac{\frac{1}{\omega L}\cos\theta-\left(\frac{1}{\omega L}\cos\theta\right)^2}{X}}}{\omega}\times 10^{-6}\ \textit{microfarads} \qquad \text{(VII)}$$

and

$$C_1=\frac{10^{-6}}{\omega\sqrt{\frac{X}{\frac{1}{\omega L}\cos\theta}-X^2}}\ \textit{microfarads} \qquad \text{(VIII)}$$

Tube Impedance

The tube impedance was found in the following manner: A non-inductive resistance R and an ammeter A were substituted in place of C_1, C, L (see Figures 5 and 7) and with a constant input I thru coil L the output determined as R was varied. This was done for several values of plate voltage. Figure 8 shows the results obtained with a 2″×2″ tube. At 700 volts and above there is a maximum at 400 ohms. Below this value the impedance of the tube seems to be higher and less well marked.

Actual test showed that within the experimental error the values of C and C_1 found agree with those calculated from equations VII and VIII when the value of tube impedance was used as determined from Figure 8.

Control Coils

Several types of control coils were wound up and compared. The results can be briefly summarized as follows:

(a) Drop turn or bank wound coils gave about 20 percent more amplification than ordinary multiple layer coils.

(b) Litzendraht was about 50 percent better than solid copper wire of comparable conductivity.

(c) The winding length should be at least 1.5 or 2 times the length of the anode.

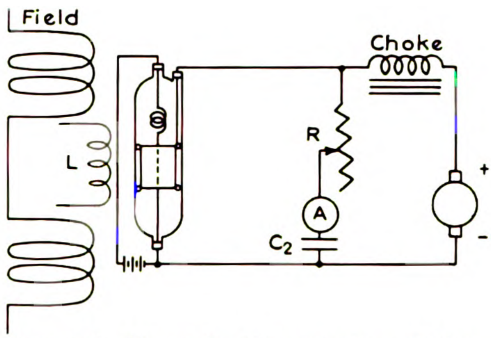

Figure 7—Circuit for Determination of Plate Resistance

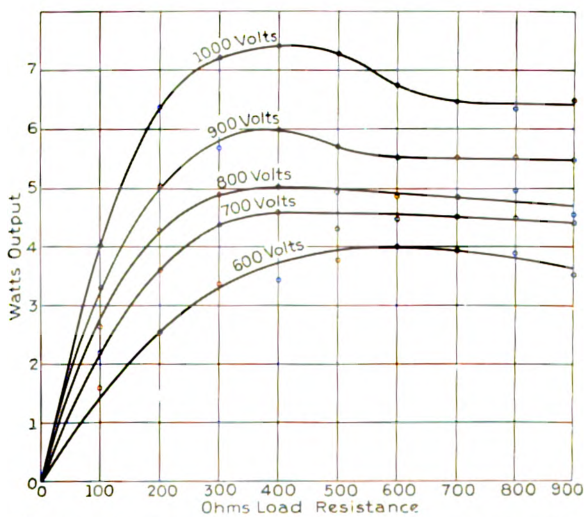

Figure 8—Output—Load Resistance Curves 2″×2″ Magnetron

Amplification

In order to determine the amplification obtainable and its

dependence upon plate voltage the apparatus was set up as indicated in Figure 9. (The field coils are not shown). Tube 2 was the tube being measured, tube 3 served as the measuring instrument. The ammeter A_2 in series with the proper resistance R, was calibrated in terms of current in coil L_2 and the plate voltage of tube 3 held constant. In order to avoid uncertainty of input resistance the input was introduced thru tube 1. The plate voltage of tube 1 was also kept constant. The amplification ratio was then the result obtained by dividing the current in L_2 as determined by tube 3, by the input current read on A_1. The results obtained are shown in Figure 10 where amplification is plotted against plate voltage. As can readily be seen the amplification is practically linear with plate voltage. It is also seen that the 1¼″×1¾″ tube gives higher amplification than the 2″×2″ tube.

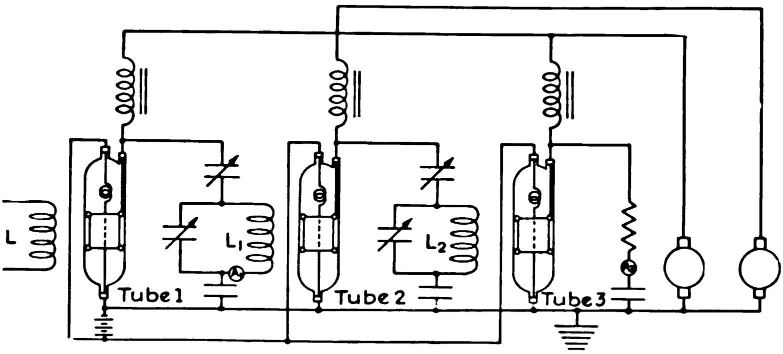

FIGURE 9—Circuit for Determination of Amplification per Tube

This method of determining the amplification has the advantage that the result found is the real amplification as obtained in practice since all losses are included. In order to be certain that these results held for the much weaker signals of radio telegraphy, the amplification was checked, using a very weak signal and testing with a small tube detector, using telephones and audibility meter or "mile-box." The results obtained by this method checked the values of Figure 10.

POLARIZING FIELD

For demonstration work a pair of Helmholtz coils of any convenient number of turns is very satisfactory, as the field can be quickly and easily adjusted; but when the building of a multistage amplifier is contemplated some consideration must be given

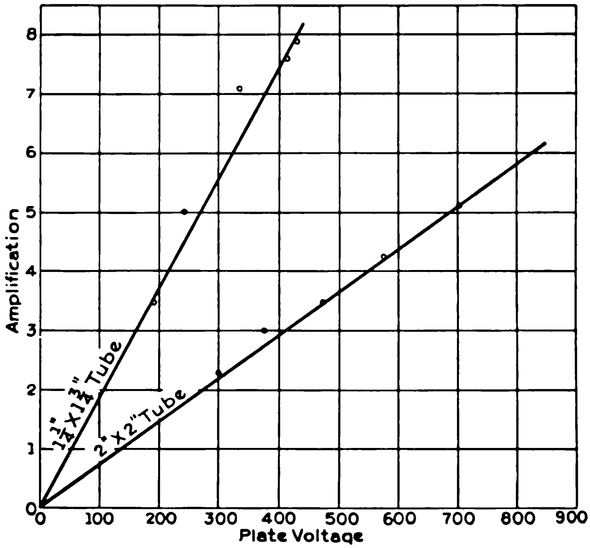

FIGURE 10—Magnetron Amplifier—Amplification per Tube

the design of field coils. In the first place the coils must be large enough so that they exercise no deleterious effect on the control coil. If the field is produced by the same voltage as that which supplies the anode of the tube, great constancy of this voltage is necessary. Figure 11 is a plot of voltage against field. The straight line represents the actual variation of field with voltage, while the parabola represents the variation required by the tube.

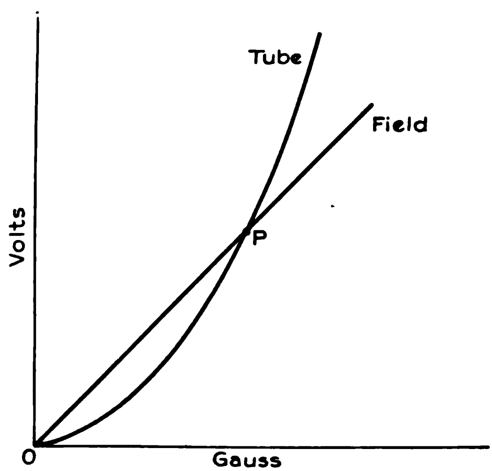

FIGURE 11—Characteristic of Magnetron with "Shunt" Field

This type of field will be called a shunt field. It is readily seen that there is only one point of intersection P. This point represents the only voltage at which the given coils produce the correct polarizing field. If, on the other hand, the field is produced by the plate current of the tube, its value will be correct at all voltages. This is shown in Figure 12. The curves are gauss-ampere characteristics of the tube at the voltages V_1, V_2, V_3; the line F represents the gauss-ampere characteristic of the field coils. While this arrangement, which will be called a series field, is always automatically stable, the field coils will be very large and heavy due to the great number of turns needed to produce the field with small plate currents, and further the resistance drop will be high.

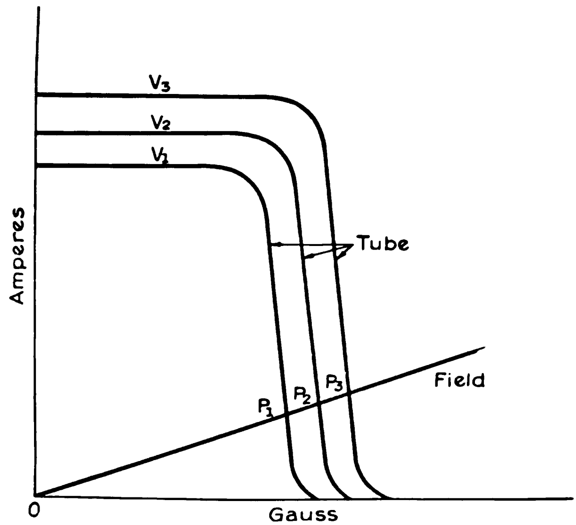

FIGURE 12—Characteristic of Magnetron with "Series" Field

It was decided, therefore, to use a combination of the two types of field, the connections being so made that the series field added to the shunt field. By adjustment of the resistance in the shunt field circuit the tube was brought to the desired operating point. It was found that when the series field was of the order of 10 percent of the shunt field the anode current remained practically constant when the anode voltage varied thru the range 400±25 volts. It was thought that this provided ample margin to care for all ordinary fluctuations of the voltage supply. It was also found that the series field was a sufficiently good choke at the frequencies used so that an additional choke coil was unnecessary.

Multi-stage Amplifier

The separate parts having been worked out, they were finally incorporated into a four-stage amplifier. The tubes used were the $1\frac{1}{4}'' \times 1\frac{3}{4}''$ tubes and the plate voltage supply was about 400 volts. This should give, according to Figure 10, an amplification of between 5- and 6-fold per stage. The control coils were wound up of litzendraht (32 strands of 0.005″ enameled wire) in the form of 10-layer bank wound coils of about 700 turns. The inductance was approximately 8 millihenrys and the resistance 6 to 6.5 ohms at a frequency corresponding to 10,000 meters. While these 700-turn 10-layer coils gave no better control than 300-turn 4-layer coils, the much higher value of inductance determined the choice, since the condenser units required were much smaller with the large coils. The diagram of connections is shown in Figure 13. As first set up, no shielding was provided and as a result never more than two stages could be operated. The amplifier was therefore rebuilt providing a copper lining to the box and a copper partition between each stage. There resulted an amplifier which was much more stable. With very careful adjustment all four stages were operable and three stages under practically any circumstances. With strong signals, such as those from the New Brunswick Station, the results were very good. When the atmospheric conditions were not too bad, signals from POZ, Nauen, Germany, were heard. The development of the magnetron amplifier was not pushed any further than this, but work taken up on the magnetron oscillator which will be the subject of the remainder of the paper.

Experience gained with the magnetron oscillator, where a

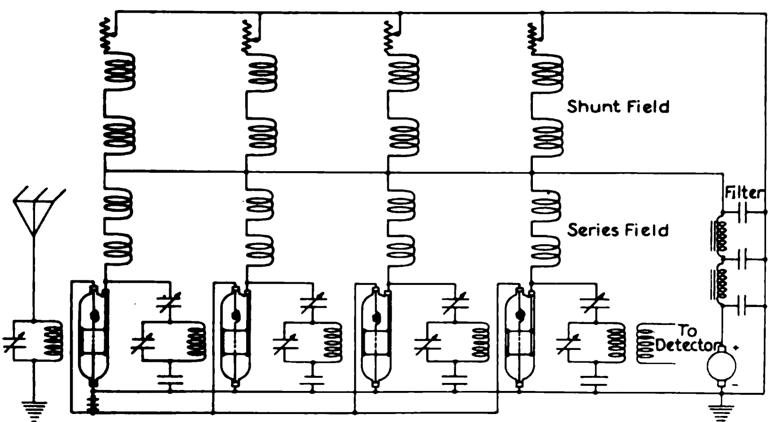

Figure 13—Circuit for Four-Stage Magnetron Amplifier

similar circuit was used, makes it seem probable that a great deal of the difficulty with the amplifier was due to improper shielding and that it would be entirely possible to redesign the amplifier and make it just as stable and give as much amplification as the same number of three-element tubes.

PART II

THE MAGNETRON AS A POWER OSCILLATOR

In the development of the magnetron oscillator, two circuits have been used. There is no choice between the two on the basis of efficiency. The first circuit was used by Mr. Glenn Mercer in the early work on the magnetron and is illsutrated in Figure 14.

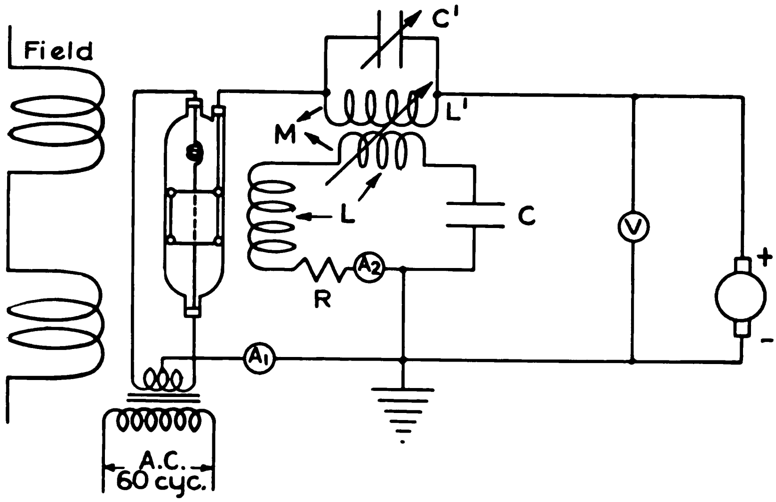

FIGURE 14—Circuit for Magnetron Oscillator

The second circuit, which has been used by the present writer, is shown in Figure 15 and will be easily recognized as the same circuit used in the amplifier but modified to suit the oscillator case. This circuit will be the only one discussed in detail. It is somewhat simpler of adjustment and in ease of operation and has been chosen for this reason.

Since the magnetron is a rather unfamiliar object outside of this laboratory, it seems advisable to give a rather full description of the component parts of the circuit in order to give a clearer conception of the proportions of these parts. For the sake of simplicity, the whole discussion will be based on the 4″×12″ magnetron. Obviously *mutatis mutandis*, it applies equally well to all sizes of tubes.

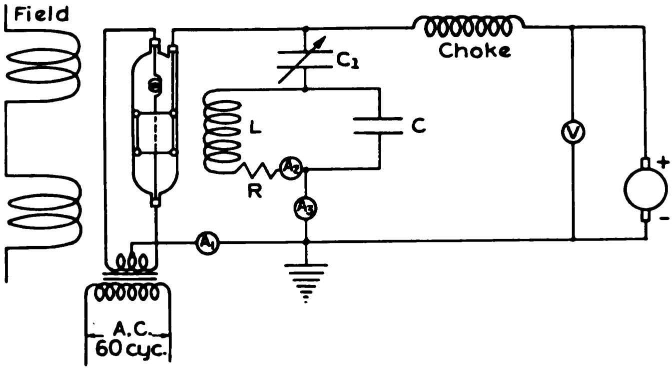

FIGURE 15—Circuit for Magnetron Oscillator

DESCRIPTION OF CIRCUIT

The "polarizing" or "biasing" magnetic field H_o is produced by a pair of Helmholtz coils or other combination of coils giving a sufficiently extended and uniform field. These coils must be large enough in diameter to avoid seriously affecting the operation of the tube by coupling with the control coil L. This coupling manifests itself in two ways: high voltages may be induced in the field coils causing breakdown of the insulation, and secondly, the apparent inductance of the coil L is decreased and its resistance increased which means a loss of useful power. This coupling effect is absent or negligible at power frequencies but becomes of more and more importance as the frequency rises.

Coils of an inside diameter of about 24″ are satisfactory. The particular set used consists of five identical coils arranged as indicated in Figure 16. Each coil has a winding section of 2″×4″, an inside diameter of 22″ wound with approximately 4,000 turns of 0.040″ enamelled wire and then impregnated with insulating varnish. Each coil weighs about 125 pounds and has a resistance of about 172 ohms. The coils are all connected in multiple across the 250-volt shop lines with a variable resistance in series. The current thru the middle coil (coil 3) was adjusted by a series resistance X, so that there resulted a field uniform to within less than 1 percent thruout a volume represented by a cylinder 6″ in diameter and 16″ long. This was done by trial using a small magnetron as an exploring instrument.[6] The constant for the system is 45.3 gauss per ampere combined current and was determined by comparison with a standard solenoid.

The filament is generally lighted from an alternating current supply. A small transformer of the proper rating with a mid-

point tap brought out from the low voltage winding has proved satisfactory.

The control coil L when used at high frequency should be wound with litzendraht in a single layer. The coil used had the following constants: 124 turns of 12-conductor litzendraht (12 insulated conductors wrapped around a rope core and again insulated, each conductor consisting of 32 strands of 0.005″ enamelled copper wire braided together). This cable will carry about 40 amperes at 30,000 cycles without too great loss. The coil was wound on a form of small maple slats giving a mean winding diameter of 7½″ and a length of 24″. The inside diameter of the slat form was 5⅞″ giving a ventilation space between it and the tube. The inductance of the coil was 0.795 millihenry and had an effective resistance of approximately 0.6 ohm measured at 12,300 meters and with a tube inside the coil.

For power frequencies the control coil presents much less of a problem. Any suitable size of wire may be used and further, the voltage per turn being less, multiple layers may be used as insulation difficulties are not encountered.

The main tuning condenser C consisted of a bank of four 0.012 microfarad mica condensers connected in multiple. The combined capacity was 0.505 microfarad, rated for 15,000 volts and safely carried 50 amperes at 10,000 to 12,000 meters.

The condenser C_1, or bridging condenser, which also serves as a stopping condenser, was variable to allow operating adjustments. The capacity range was 0.0013 to 0.0029 microfarad and would withstand 20,000 volts.

The load is represented by the resistance R. The total resistance of the circuit was of the order of 6 ohms.

The choke coil was a single layer solenoid wound with 0.064″ double cotton-covered wire and treated with insulating varnish. The form was a piece of insulating tubing 13¼″ in diameter by 48″ long. The maximum inductance was 30.5 millihenry and convenient taps were brought out to allow adjustment.

Tube Operation

It is obvious that for efficient operation of the tube, the plate current must flow at a time when the plate voltage is low. The control field must therefore be in phase with the instantaneous plate voltage. The instantaneous plate current which flows will thus be 180° out of phase with plate voltage and control field. Such a set of conditions is represented in Figure 8. At (a) is shown the instantaneous plate voltage e_p as a sine wave of maxi-

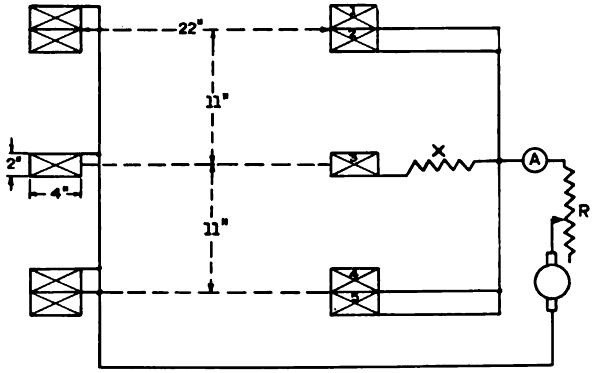

FIGURE 16—Magnetic Field—Coil System

mum amplitude V superimposed on the impressed direct voltage V_o. The minimum value of e_p will be termed the emission voltage V_e.

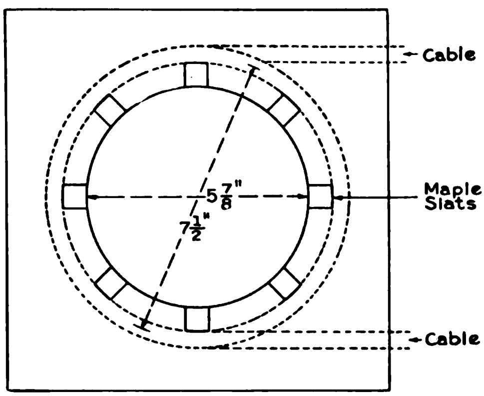

FIGURE 17—Section Thru Control Coil Support

At (*b*) is shown the instantaneous plate current i_p, trapezoidal in shape, rising from a value zero to a value I_e in the time angle $(\theta_2-\theta_1)$, remaining equal to I_e for the time 2 θ_1, and falling to zero again in the time $(\theta_2-\theta_1)$. The average value of i_p as read on a direct current ammeter is shown by the dotted line I_{pa}.

Finally at (*c*) is shown the control magnetic field as a sine wave of amplitude H superimposed on the "biasing" field. H_o.

While at first sight the choice of wave shapes seems rather arbitrary, nevertheless such is not the case. As the alternating control field is of considerable magnitude, the current thru the coil must be large, and thus there is a large ratio of circulating

volt-amperes to dissipated watts. The control current is thus very closely sinusoidal and produces a sinusoidal voltage across the coil L and condenser C. The plate voltage e_p also will be nearly sinusoidal in the absence of other means of distorting it, and oscillograms show this to be true. The choice of a trapezoidal shape for the plate current rests entirely on experiment, that is, the oscillographic record. Measurements on films show the time angle $(\theta_2-\theta_1)$, to be approximately $\frac{\pi}{6}$ and for simplicity of calculation has been so chosen. The departures from these simple forms in actual practice are slight and lead to no appreciable errors.

Having considered the wave forms, the actual magnitudes can be discussed in a general way. The tube is characterized by three important features: (a) voltage rating, (b) filament rating, and (c) the maximum safe continuous power dissipation at the anode. The voltage rating determines the applied direct voltage V_o, while the filament rating fixes the maximum current that should be drawn thru the tube. This value of current has been designated I_e and is the maximum electron emission compatible with the desired life of the filament. The power dissipation of the anode determines the maximum output permissible.

Efficient use of the electron emission from the filament then demands that the minimum plate voltage does not fall too low. By use of equation (1) the proper value of V_e can be calculated.

The values of H_o and H will be determined by applied plate voltage V_o, the constants of the control coil L, and the power output desired. A rather wide variation of relative values of H_o and H is permissible, thus allowing considerable latitude for adapting them to the case at hand. In any case, however, the value of H should not be much in excess of the value of H_o, since then there is a possibility of a field (in the reverse direction to H_o) of sufficient strength to prevent electrons from reaching the anode at the voltage V_e.

Having thus considered the general features of tube operation, detailed calculation of expected performance can be made. Referring to Figure 18, let,

V_o = impressed direct voltage measured with respect to the filament.

V_e = minimum instantaneous plate voltage.

V = amplitude of alternating plate voltage.

e_p = instantaneous plate voltage with respect to the filament.

i_p = instantaneous plate current.
I_e = maximum instantaneous plate current.
I_{pa} = average plate current.
H_o = "polarizing" or "biasing" field.
H = amplitude of alternating control field.
θ_i = time angle during which current flows.

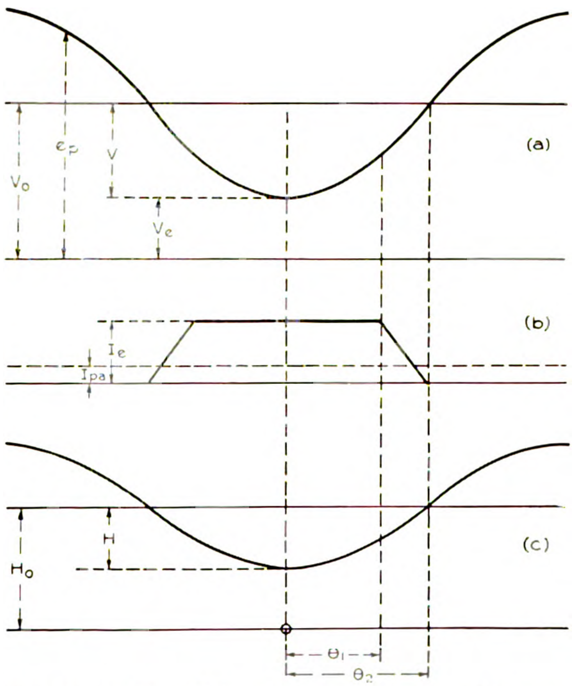

FIGURE 18—Plate Voltage, Plate Current and Magnetic Field Relationships in Magnetron Oscillator

It then follows that neglecting the filament loss, the instantaneous power input into the system is $V_o i_p$, the instantaneous tube loss is $e_p i_p$, the instantaneous output is $V_o i_p - e_p i_p$, the instantaneous efficiency is $\frac{V_o i_p - e_p i_p}{V_o i_p}$. In order to obtain average values, these instantaneous values must be integrated over a complete cycle, thus the

Average power input is $$P = \frac{1}{2\pi}\int_o^{2\pi} V_o i_p \, d\theta = V_o I_{pa} \tag{IX}$$

Average tube loss is $$P_1 = \frac{1}{2\pi}\int_o^{2\pi} e_p i_p \, d\theta \tag{X}$$

Average tube output is $P-P_1=V_o I_{pa}-\frac{1}{2\pi}\int_o^{2\pi} e_p i_p d\theta$ (XI)

Average tube efficiency is $\eta=\dfrac{V_o I_{pa}-\frac{1}{2\pi}\int_o^{2\pi} e_p i_p d\theta}{V_o I_{pa}}$ (XII)

If at the time $t=0$, $\theta=0$, and $e_p=V_e$, then we have $e_p=V_o-V\cos\theta$, and $i_p=I_e$ from $\theta=0$ to $\theta=\theta_1$, and $i_p=I_e\left(\frac{\theta_2-\theta}{\theta_2-\theta_1}\right)$ from $\theta=\theta_1$ to $\theta=\theta_2$.

Since the cycle is symmetrical about 0, and no current flows after $\theta=\theta_2$, we have for the average plate current $I_{pa}=\frac{I_e(\theta_2+\theta_1)}{2\pi}$ and for the average tube loss $P_1=\frac{1}{2\pi}\int_{-\theta_2}^{\theta_2} e_p i_p d\theta$ or

$P_1=\frac{1}{\pi}\int_o^{\theta_2} e_p i_p d\theta.$

Substituting in the value of e_p and i_p

$$P_1=\frac{1}{\pi}\int_o^{\theta_1} I_e(V_o-V\cos\theta)\,d\theta+\frac{1}{\pi}\int_{\theta_1}^{\theta_2} i_e\left(\frac{\theta_2-\theta}{\theta_2-\theta_1}\right)(V_o-V\cos\theta)\,d\theta$$

and carrying out the integrations indicated,

$$P_1=\frac{I_e V_o(\theta_2-\theta_1)}{2\pi}+\frac{I_e V}{\pi(\theta_2-\theta_1)}(\cos\theta_2-\cos\theta_1)$$

Now the power input is $P=V_o I_{pa}=I_e V_o\left(\frac{\theta_2+\theta_1}{2\pi}\right)$, and hence the power output is

$$P_2=-\frac{I_e V}{\pi(\theta_2-\theta_1)}(\cos\theta_2-\cos\theta_1)=\frac{I_e V}{\pi(\theta_2-\theta_1)}(\cos\theta_1-\cos\theta_2)$$

and the efficiency is $\eta=\frac{V}{V_o}(\cos\theta_1-\cos\theta_2)\dfrac{1}{(\theta_2-\theta_1)\left(\frac{\theta_2+\theta_1}{2}\right)}$.

Experiment having shown that $\theta_2-\theta_1$ is $\frac{\pi}{6}$ there finally results

$$P=\frac{I_e V_o}{2\pi}\left(2\theta_1+\frac{\pi}{6}\right) \quad \text{(XIII)}$$

$$P_1=\frac{I_e V_o}{2\pi}\left(2\theta_1+\frac{\pi}{6}\right)-\frac{6}{\pi^2}I_e V(\cos\theta_1-\cos\theta_2) \quad \text{(XIV)}$$

$$P_2=\frac{6}{\pi^2}I_e V(\cos\theta_1-\cos\theta_2) \quad \text{(XV)}$$

$$\eta=\frac{6}{\pi}\left(\frac{V}{V_o}\right)(\cos\theta_1-\cos\theta_2)\frac{1}{\left(2\theta_1+\frac{\pi}{6}\right)} \quad \text{(XVI)}$$

In Table I are given data for use in equations XIII to XVI calculated for values of θ_1 from 0° to 90°.

TABLE I

$\theta_1{}^\circ$	$\cos\theta_1$	$\cos\theta_1 - \cos\theta_2$ $\theta_2=\theta_1+30^\circ$	$(\cos\theta_1 - \cos\theta_2)\frac{6}{\pi^2}$	$\frac{\theta_1+\theta_2}{2\pi}$	$\frac{(\cos\theta_1 - \cos\theta_2)\frac{6}{\pi^2}}{\frac{\theta_1+\theta_2}{2\pi}}$
0	1.000	0.1340	0.0815	0.0833	0.9779
5	0.9962	0.1770	0.1076	0.1111	0.9685
10	0.9848	0.2188	0.1330	0.1389	0.9576
15	0.9659	0.2588	0.1573	0.1667	0.9438
20	0.9397	0.2969	0.1805	0.1944	0.9285
25	0.9063	0.3327	0.2023	0.2222	0.9102
30	0.8660	0.3660	0.2225	0.2500	0.8900
35	0.8192	0.3966	0.2411	0.2778	0.8679
40	0.7660	0.4240	0.2578	0.3056	0.8435
45	0.7071	0.4483	0.2725	0.3333	0.8177
50	0.6428	0.4692	0.2852	0.3611	0.7899
55	0.5736	0.4864	0.2957	0.3889	0.7603
60	0.5000	0.5000	0.3040	0.4167	0.7295
65	0.4226	0.5098	0.3099	0.4444	0.6974
70	0.3420	0.5156	0.3134	0.4722	0.6638
75	0.2588	0.5176	0.3147	0.5000	0.6293
80	0.1736	0.5156	0.3134	0.5278	0.5939
85	0.0872	0.5098	0.3099	0.5555	0.5579
90	0.0000	0.5000	0.3040	0.5833	0.5211

The relation between plate voltage and plate current is given in Figure 19 as calculated from equation (I). The ratio of maximum plate voltage swing to impressed voltage or $\frac{V}{V_o}$ is given in Figure 20 for values of emission current I_e from 0 to 10 amperes and from 5,000, 10,000, and 15,000 volts for V_o.

Figure 21 gives the average plate current I_{pa} as a function of time angle θ for emission values of 1, 2, 4, 6, 8, and 10 amperes.

A set of input, output, loss, and efficiency curves has been calculated from the data of Table I and figures 19, 20, and 21, for filament emissions of 2, 4, 6, and 8 amperes and an applied direct voltage of 10,000 volts. These curves are plotted in Figure 22. Inspection of this figure shows that maximum output is obtained for a time angle of $\theta_1 = 75^\circ$ and, based on this value, Figure 23 represents the relation between output and applied voltage for a filament emission of 4 amperes.

Circuit Design

In order that the calculated performance may be realized the various parts of the circuit must be properly proportioned. In Figure 24 is shown the vector diagram of the current and voltage relationships of the circuit, tho somewhat distorted in order to obtain clearness in the drawing. This diagram is drawn keeping in mind the fundamental facts (see Figure 18) that plate voltage

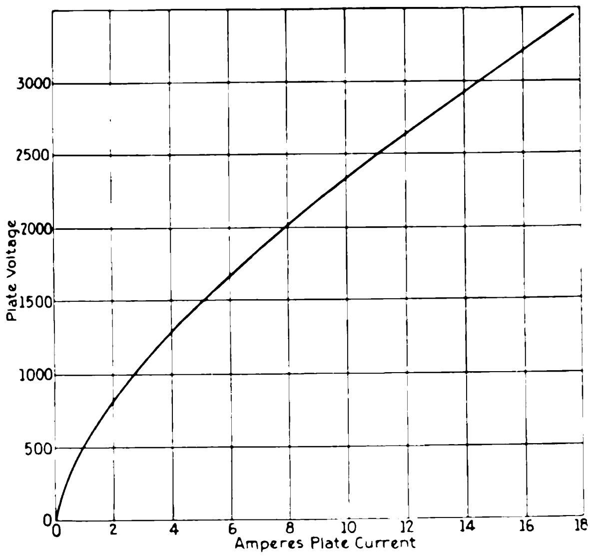

Figure 19—Space Charge Current for 4″×12″ Magnetron

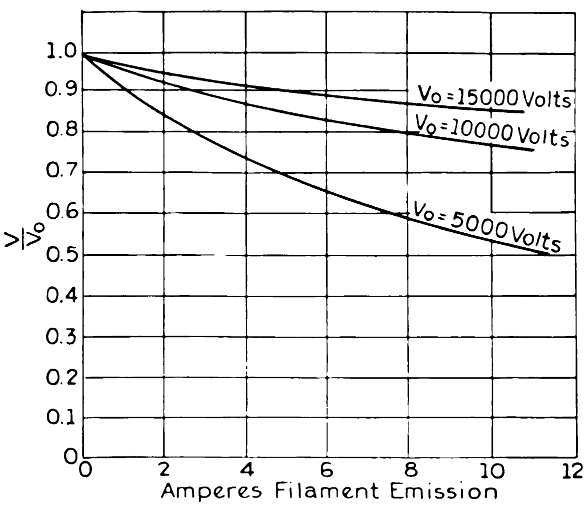

Figure 20

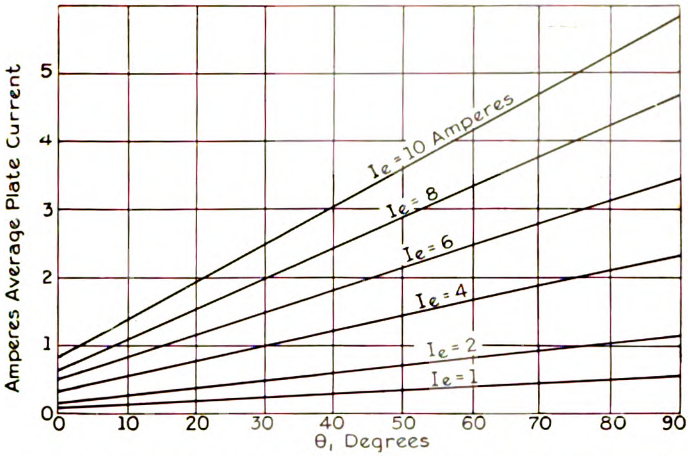

FIGURE 21—Average Plate Current as a Function of Time Angle 4″×12″ Magnetron

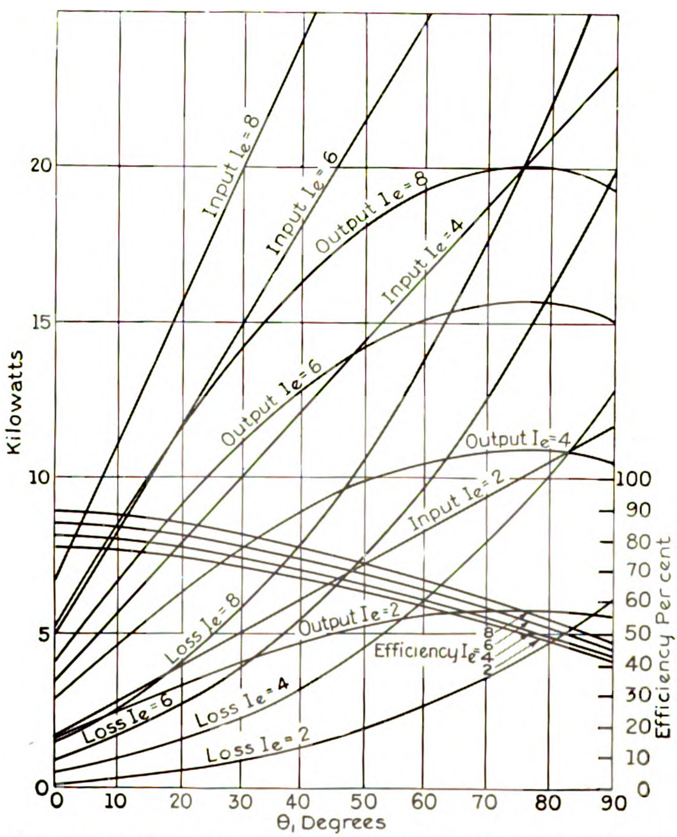

FIGURE 22—Input, Output, Loss and Efficiency Curves for 4″×12″ Magnetron Operating at 10,000 volts, D.C.

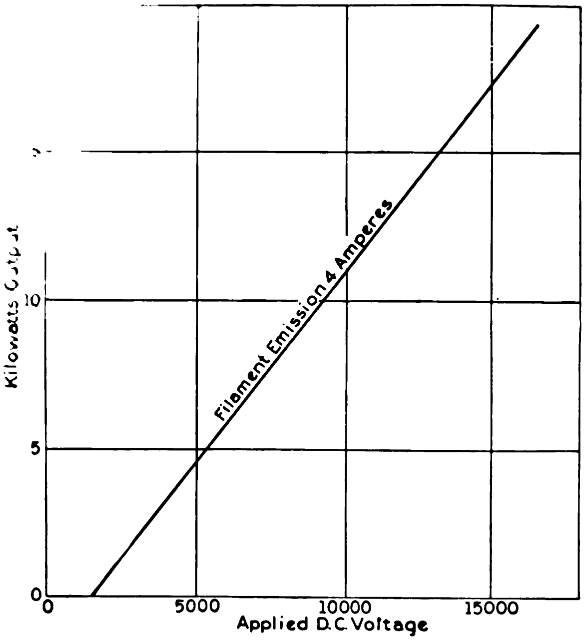

Figure 23—Output as a Function of Plate Voltage for 4″×12″ Magnetron

should be in phase with control current and that plate current should be 180° out of phase with plate voltage, therefore I_L and e_p are laid off in the same direction and to the proper scale, while i_p is laid off in the opposite direction along the same line. The current I_L flows thru inductance L and resistance R which gives a component $\omega L I_L$ lagging 90° behind I_L and a component $R I_L$ in phase with I_L (see Figure 15). The resultant voltage E_c is also impressed across condenser C and thus gives rise to the current I_c leading E_c by 90°. As a resultant of I_L and I_c we have the series line current I_{c_1}, and this current flowing thru condenser C_1 gives the voltage E_{c_1} where I_{c_1} leads E_{c_1} by 90°. The plate voltage e_p must then be the resultant of E_c and E_{c_1}. If it were not for the choke coil the plate current would have to be I_c', which is I_{c_1} reversed, since the vector sum of the currents leaving and entering such a point as 0 must be zero. The plate voltage e_p is also impressed across the choke coil, thus e_{choke} coincides with e_p, but the ckoke draws a current I_{choke} lagging 90° behind e_{choke}. This choke current then combines with I_{c_1}' to give as a resultant i_p and if the choke is properly designed i_p can be brought in line with I_L and e_p and in the opposite direction, as was required initially.

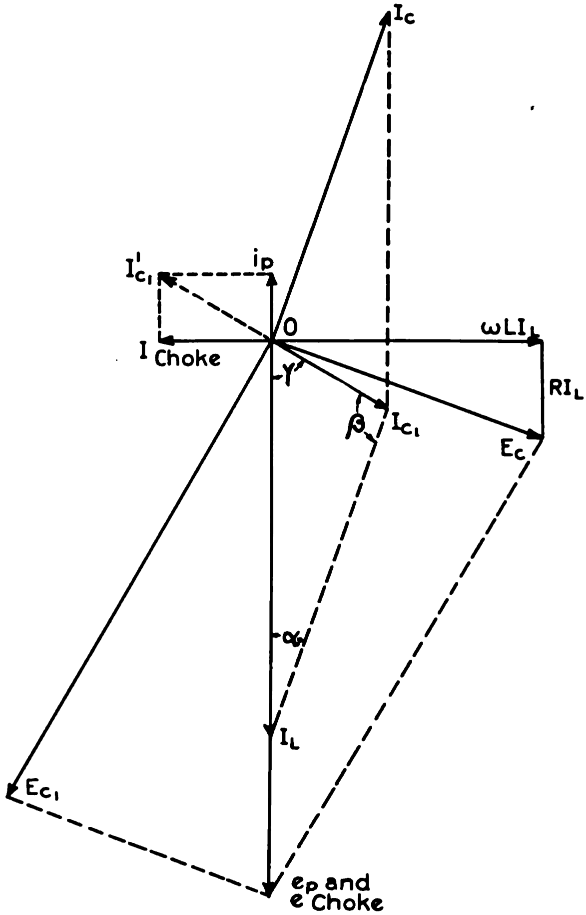

FIGURE 24—Vector Diagram for Magnetron Oscillator

A further consideration of this diagram will show that the frequency of oscillation ω will be such that $\omega < \omega_o$ where ω_o corresponds to the frequency calculated from $\omega_o^2 L C = 1$.

One further point must be considered in circuit design and that is the equivalent series resistance of the circuit L, C, R. The tube cannot operate efficiently unless this circuit has the proper resistance value. The curves shown in Figure 25 were taken with a 4″×12″ tube working into a pure resistance (see Figure 7). As can be readily seen, the curves are relatively flat but show a slight maximum at about 1,500 ohms. Numerous tests have shown that best results are obtained when $\frac{I_L^2 R}{I_{c_1}^2} = 1{,}500$ ap-approximately.

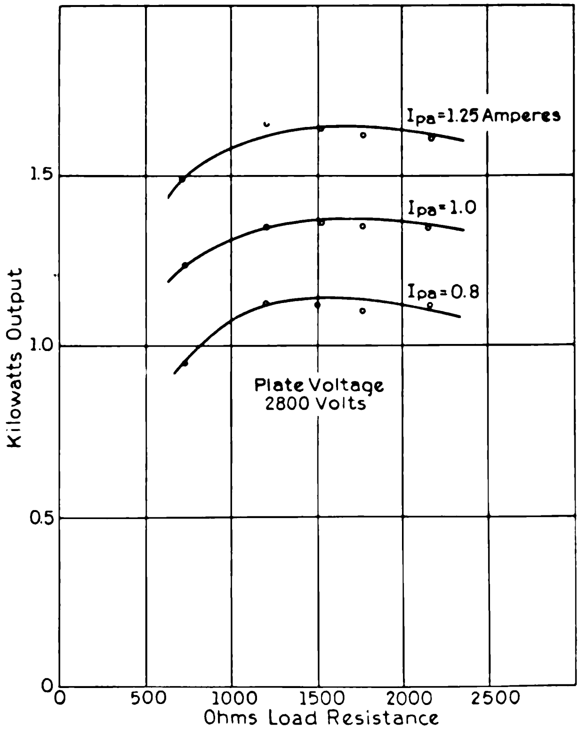

FIGURE 25—Plate Resistance of 4″×12″ Magnetron

As an example the detailed calculation of a circuit will be given An output of 10 kw. is desired at a frequency of 15,000 meters and the available direct voltage is 10,000 volts.

The tube is a 4″×12″ magnetron and has a filament rating of 22 volts and 17.5 amperes. (This corresponds to an 0.020″ diameter tungsten filament running at 2,500° K., and will give an emission of 4 amperes with a life of about 1,000 hours.) The safe continuous plate dissipation is 5 kw.

The control coil L will be assumed to have an inductance of 1.0 millihenry and wound 2 turns per cm. with cable suitable to carry 40 amperes at this frequency.

Referring now to Figure 22 on the curve for $I_e = 4$ amperes, it is found that the angle $\theta_1 = 50°$ corresponds to an output of 10 kw. and a loss of 4.45 kw. The filament loss is 0.385 kw., giving a total loss of 4.84 kw., which is well within the tube rating. From Figure 21 we find that the average plate current is 1.445 amperes. Since the plate supply is 10,000 volts, the total plate input is 14.45 kw., which checks the loss output figures.

Since the wire of the control coil is rated at 40 amperes, we must have $R = 6.25$ ohms. The $R\,I_L$ drop then becomes 250 volts, and since 15,000 meters corresponds to $\omega = 125{,}000$, we have $\omega\,L\,I_L = 5{,}000$ volts. This gives $E_c = 5{,}006$ volts. The angle a is given by $tan\,a = \dfrac{R}{\omega\,L}$ or $tan\,a = 0.05$ and $a = 2°\ 51.'7$.

Since the equivalent series resistance of the circuit must be 1,500 ohms, we have $I_{c_1}{}^2\,R' = I_L{}^2R$ from which $I_{c1} = 2.582$ amperes

By the sine rule $\qquad \dfrac{I_c}{sin\,a} = \dfrac{I_L}{sin\,\beta} = \dfrac{I_c}{sin\,\gamma}$

and applying this $\beta = 129°\,19'$, $\gamma = 47°\,59'.3$, and $I_c = 38.315$ amperes. Since $I_c = C\,\omega\,E_c$, we have $C = 0.06123$ mf.

As a check value on these values, we may calculate the equivalent series resistance of the L, C, R circuit from the formula

$$R' = \frac{R}{m^2\,R^2\dfrac{C}{L} + (m^2 - 1)^2}$$

where R is the actual resistance in the circuit, L the inductance, C the capacity, and m the ratio of the actual frequency to the resonant frequency.

In this case $R = 6.25$ ohms, $C = 0.06123$ mf., $L = 1.0$ millihenry, and $\omega_o = 127{,}800$ since $\omega_o\sqrt{LC} = 1$. This gives $m = 0.9781$ and $m^2\,R^2\dfrac{C}{L} + (m^2 - 1)^2 = 0.004163$ from which $R' = 1{,}501.3$ ohms.

For a current of 4 amperes thru the tube, Figure 10 shows the necessary voltage to be 1,300 volts, which is then the value of V_e. This gives $V = 8{,}700$ volts, but as the vector diagram deals with root-mean-square values we have for e_p the value 6,152 volts.

In Figure 24 it is seen that $E_{c1} = \sqrt{(\omega\,L\,I_L)^2 + (e_p - R\,I_L)^2}$. Now $\omega\,L\,I_L = 5{,}000$ volts and $e_p - R\,I_L = 5{,}902$ volts, which gives $E_{c_1} = 7{,}735$ volts. This value E_{c_1} must be produced by the current I_{c_1} flowing thru C_1, hence $I_{c_1} = C_1\,\omega\,E_{c_1}$ from which $C_1 = 0.00267$ mf.

For the choke coil current we have $I_{choke} = I_{c_1}\,sin\,\gamma$ or $I_{choke} = 1.914$ amperes. The voltage across the choke coil being e_p in value or 6,152 volts, the choke coil must then have an inductance of 25.72 millihenrys.

There finally remains the calculation of the polarizing field necessary. The angle θ_1 has been determined as 50°, so that $\theta_2° = 80°$. The magnetic field must therefore become of such a value as to prevent electrons from reaching the anode at this time. The plate voltage is given by $V_o - V\,cos\,80°$, which is 8,500

volts. Referring now to Figure 26, it is seen that the magnetic field necessary at this voltage is 121 gausses. Using the formula $H = 0.4\pi NI$, the current constant of the control coil is found to be 2.5 gausses per ampere, and as $I_L = 40$ amperes (root-mean-square), $H = 141$ gausses. At the time $\theta_2 = 80°$, $H \cos\theta_2 = 24.5$ gausses. This gives then for the polarizing field $H_o = 145.5$ gausses.

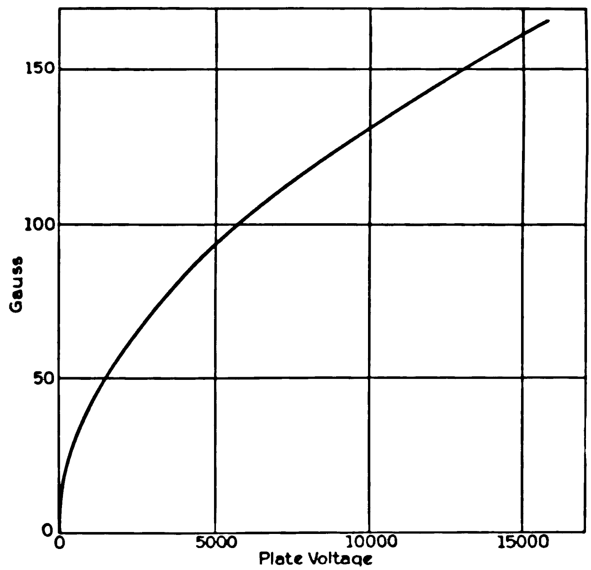

FIGURE 26—Critical Field for 4″×12″ Magnetron.
$V = 0.0221\, r^2 H^2$

By proper design of the coils the field can be conveniently furnished by the average plate current. In order to do this, the filament transformer mid-tap is connected to one terminal of the coil system and the other terminal grounded. It is then necessary to shunt the coils with sufficient capacity to by-pass the alternating component of the plate current. If sufficient copper is used in the coils the RI drops will be low and the loss correspondingly small. This scheme has the advantage of automatically adjusting itself, but it suffers from the lack of flexibility.

Some form of forced ventilation between the control coil and the glass tube is necessary in the large sizes. In order that a 4″×12″ tube may dissipate 6 kw. continuously at the anode, a half horsepower blower furnishes sufficient air.

Tests

Numerous tests have been carried out at both high and low frequency, and several oscillograph films taken. In Table II will be found a few typical results. These tests were made at different times and with different tubes. The wave length was about 12,000 meters, and the tubes were self-excited. Under the heading, "calculated efficiency," the calculations were based on a value of 4 amperes for filament emission since the filaments were run at a temperature such as to give this value.

TABLE II

V_o Volts	I_{pa} Amps.	Input Kw.	Output Kw.	Loss Kw.	Efficiency Percent	Calculated Efficiency Percent
3,000	1.0	3.0	1.55	1.45	52	51
5,000	1.0	5.0	3.25	1.75	65	65
5,000	1.45	7.25	4.45	2.80	61.3	56.8
7,000	1.1	7.7	5.10	2.60	66.2	66.6
7,500	1.4	10.5	6.80	3.70	64.8	67.5
9,000	1.3	11.7	8.05	3.65	68.8	69.3

As will be readily seen, the results obtained agree as well as can be expected with the calculated performance. In order to have a permanent record a few films were made at lower frequency. Figure 27 was taken with the tube non-oscillating, that is, with the tube directly across the generator, and there was no load on the tube. Curve *A* is the 40-cycle control current of 6.2 amperes thru a coil giving 8.8 gausses per ampere. The maximum control field is thus 77.5 gausses. The permanent polarizing field was 60 gausses. Curve *B* shows the current thru the tube. By measurement on the film, current begins to flow when the control field is at a value of 13 gausses below the zero line, but as this zero line corresponds to 60 gausses the true field is 47 gausses. The theoretical value of field is found to be 46 gausses (see Figure 26) for 1,200 volts, which was the voltage applied. In curve *C* is shown the generator voltage. The oscillatory character of the voltage is explained by fact that a capacity of a few microfarads was placed across the generator terminals. The object of this

was to hold the voltage constant and prevent damage to the generator due to the transient produced by the sudden starting and stopping of the plate current. Since the generator armature has inductance, the capacity formed a resonant circuit with this inductance and shock excitation caused the oscillations. On measurement it will be found that the plate current in curve *B* rises to a maximum in a time angle of considerably less than 30°, but it must be remembered that there was no load on the tube. The effect of the resistance of the working circuit is to make this rise in current take a longer time.

Figure 28 is from a film taken with the tube oscillating at 500 cycles. For convenience of measurement the tube was excited from a 500-cycle generator and curve A shows this control current. The zero line corresponds to a polarizing field of 100

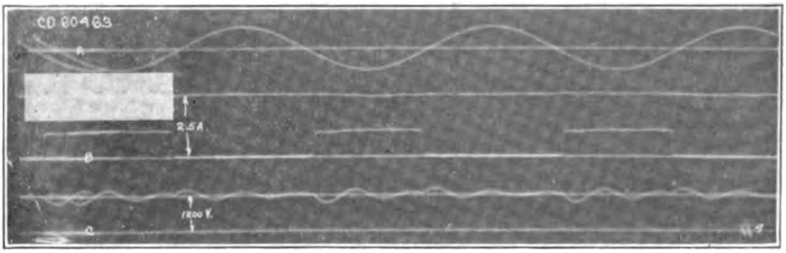

FIGURE 27—Characteristic of 4"×12" Magnetron, Non-oscillating, Taken by Oscillograph

gausses. The control current of 23 amperes thru a coil giving 2.68 gausses per ampere gives a peak value of 87 gausses. Curve *B* is the plate current rising to a maximum of approximately 3 amperes in a time angle of about 30°, remaining practically constant and then falling to zero again in the same angle. Curve *C* shows the plate voltage which will be noted to be nearly a sine wave. The calibration line corresponds to 2,600 volts. Measurement on the film shows that the tube

	Volts	Gausses	Gausses (Calc. *H* from Figure 26)
begins to open at........	2,170	61	61.5
is entirely open at.......	1,700	33.5	54.5
begins to close at........	1,950	59	59.5
is entirely closed at......	2,250	94	62.5

It is important to notice that the plate current begins to flow and starts to stop at the time when the field and voltage arrive simultaneously at the proper values. The adjustment of the circuit was not particularly good and the plate voltage does not fall quite low enough for the proper utilization of the emission. Thus the output was low, the loss high, and the efficiency low, however the principal features are clearly brought out.

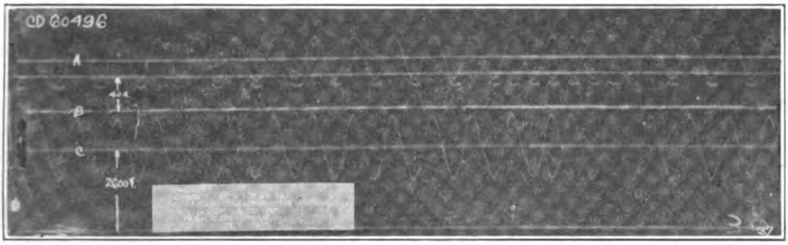

FIGURE 28—Characteristic of 4"×12" Magnetron Oscillating at 500 Cycles (approx.), Taken by Oscillograph

CONCLUSION

Space does not permit a complete discussion of all the factors which may affect tube operation. A brief mention will be made of a few rather obvious conditions which may exist and their general effect on the system. If the plate voltage remains too high while current flows thru the tube, the plate loss is increased and the output decreased. If the plate voltage falls too low, the power output is decreased, but the efficiency may increase over a limited range. If with properly adjusted values of voltage and current the phase relations become disturbed, then some plate current is drawn at too high voltage and the loss is increased. In this case the output may be maintained but the efficiency falls off.

Of course greater efficiency could be obtained by properly deforming plate voltage and plate current so as more nearly to approach "square" waves. Practical ways of producing these results are not very simple and the gain is apt to be offset by the difficulties encountered.

If the tube is separately excited, the conditions outlined in the paper can be accurately obtained. When the tube is self-excited slight departures may occur, but as shown by experiment these deviations do not seem to be sufficient to cause any appreciable error in the calculations.

There is a limitation to the magnetron at the very high frequencies, and it arises from the difficulty of producing the neces-

sary alternating magnetic field. In order to have very high frequency the inductance must be small; in other words, a very few turns of wire and hence excessively large currents are required to produce the magnetic field.

The magnetron has the advantage of very simple construction and can be made in almost any desired size and the efficiency of operation compares favorably with radiotrons of comparable output.

The writer wishes to acknowledge his indebtedness to Dr. A. W. Hull for his advice and criticism thruout the course of this work, to Mr. E. F. Hennelly for his co-operation in the design and production of tubes, and to Mr. E. W. Kellogg and Mr. D. C. Prince for helpful suggestions.

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SUMMARY: As an introduction, details of tube design are discussed, together with a partial list of the sizes studied. A resumé of the general theory of magnetic control is then given.

The first part of the paper is taken up with the application of the magnetron as an amplifier at 8,000 meters wave length. The circuit is described and the results of tests on tube impedance, design of control coils, the variation of amplification with anode voltage, design of polarizing field coils, and description of a four-stage amplifier are given.

The second part is concerned with the use of the magnetron as a generator. Since the magnetron is a rather unfamiliar tube, the circuit and necessary apparatus are described in considerable detail. The conditions necessary for efficient operation are then discussed. Based on observed wave shapes of anode voltage and anode current, formulas suitable for circuit design are developed, and the complete calculation of a typical circuit given. The results of tests at various voltages are then compared with calculated performance and representative oscillograms shown. In conclusion, a few factors which may cause departure from theoretical results are briefly discussed.

NOVEL CURRENT SUPPLY SYSTEM FOR AUDIONS*

By

Charles V. Logwood

(De Forest Radio Telephone Company, Jersey City, New Jersey)

Never in the history of mankind has invention been stimulated to such an extent as it has by the three-electrode vacuum tube invention of Dr. Lee De Forest. This invention is used in a combination of three industries, namely, the long distance telephone, radio communication, and radio broadcasting, which have grown to gigantic proportions in the last decade. The minds of thousands of trained engineers are daily applied to the task of discovering new application of the established properties of the audion. We find methods of accomplishing desired results described in many publications in dealing with its use in new combinations and in data on the circuits interlinking the input and output branches of the audion device and interlinking audions and their circuits to one another. In any case, we must deal with the power supply to the audion itself. This paper is therefore not entirely concerned with the audion alone. Engineers working with audions have from time to time developed new and important discoveries and applications of standard uses of other devices to the problem of supplying the audion with the power for heating a hot cathode as well as a power supply for the plate circuit.

Both of these power supplies must be constant in their voltage because otherwise causes independent of the controlling forces present in the grid circuits would cause fluctuations in the output circuits of the audion as, for example, when the voltage was being varied at the source supplying the power to heat the filament. This change would cause changes in the plate circuit current detrimental to successful operation. If the plate circuit supply were likewise of a type having inconstant voltage, the variations would cause a disturbance in the output circuit.

The usual source of constant power supply is a storage battery. This device imitates the properties of a condenser of very

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great capacity as well as those of a secondary generator. Altho a storage battery operates well on both filament and plate circuits, the use of most of the present type of cells to produce 90 volts in the plate circuit would result in an expensive and unwieldy storage battery for plate supply. Small batteries have been specially developed for this purpose. These batteries have been giving very satisfactory service when properly proportioned in size to meet the requirements.

Dry batteries have been improved to a very great extent and tremendous numbers of them have been made to supply the demand, but the rapid evolution and production of five or more tube sets has resulted in the drawing of heavy current loads from these batteries. This has caused general dissatisfaction to such an extent that substitutes have sprung up everywhere utilizing line power sources of alternating or direct current to feed the plate circuit of the audions. When too much current is drawn from dry batteries, cell leakage occurs, and therefore noises are heard in the telephones which are annoying to the person operating the receiving set. Dry batteries which are of an inferior make give constant trouble because of cell leakage within the battery itself. This fact introduces an obstacle to satisfactory service, and the shelf life of the battery may be short, and therefore a source of general discontent with batteries as a whole.

There are many methods that have been worked out for supplying filament current without the use of storage batteries. One of these has been before the public for the last few years. This solution is made possible by the re-design of the audion itself. The audion filament is made to consume a very small amount of power from the dry battery. This solution to my mind is not ideal, as it involves the life of the tube. The dimensions of this type of audion filament are such as to make it produce loud microphonic sounds unless cushioned to meet any shock applied to it. Such vibration in itself reduces the useful life of the audion. The manufacture of dry cell tubes is an exceedingly difficult matter. This type of audion is also difficult to handle. The tubes should never be operated at more than ten percent and preferably at about five percent of the total emission, and unless one is skilled in the art and can remain within the operating limits of the dry cell audion, the life of these tubes will vary considerably among themselves, and therefore, in practice, the life of the dry battery tube is shorter than the life of the storage battery filament tube. The filaments of both are constructed of the same material. For example, the De Forest storage battery tube

at the laboratory will operate for a period of five thousand hours; whereas the dry battery tube will operate for a period of three thousand hours. These lives are the times during which the emission drops five percent below the first stable emission.

Elimination of Ripple Due to Circuits

One of the simplest forms of ripple elimination is that of feeding the filament from an alternating current source and connecting the grid circuit to the filament at a point having a potential intermediate between the potentials of the ends of the filament. This very simple method of heating a filament from an alternating current supply has been utilized by engineers and is a fairly old one. It is shown in Figure 1.

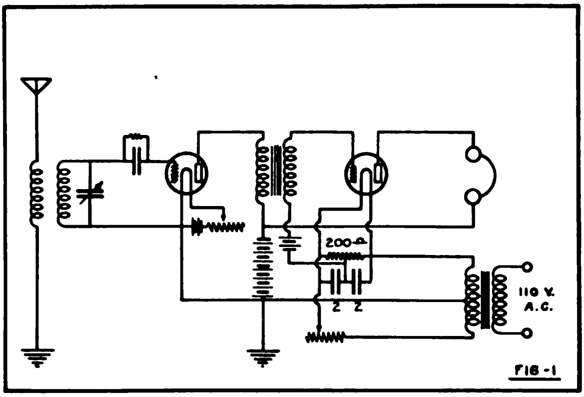


Figure 1

The Cooper-Hewitt mercury rectifier for garages and for central telephone station operation has been in use for more than ten years. Previous to the mercury valve rectifier, Mr. Thomas Edison discovered the thermionic rectifier. Its operating principle, the Edison effect, was revived by Fleming of England when he made it perform as a detector of Hertzian waves. Considerable patent litigation has resulted since Dr. De Forest inserted the grid in the Edison valve. It is rather difficult to give the credit to any one for the rectifier principle other than Edison, since he opened the way for inventors to study the resulting effects.

It did not take the experts long to find that alternating current for the filaments was a hard proposition to handle. Engineers know very well what occurs when an audion detector filament is lit with commercial alternating current.

However, under certain conditions the amplifier tubes can be lit on alternating current and yet very little hum is heard in the telephones. A simple method is shown in Figures 1 and 2.

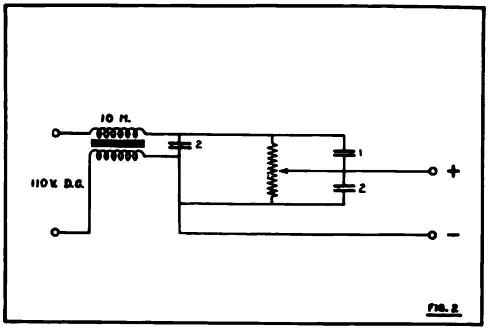


FIGURE 2

Attention is directed particularly to Figure 2. The device which is included in this circuit was supplied to the Navy Department in the year 1916.[1] The source of current was the ships mains. The filter coil was a double winding affair with an open core. One section of the winding was in the negative lead and the other in the positive lead. Across the output terminals of this filter coil was connected a three-terminal potentiometer bridged by proper condensers. The variable contact of the potentiometer leads to the receiving set. The potentiometer was made of hard graphite of about 12,000 ohms resistance. The potential delivered to the receiving set was regulated by this variable contact. The type of filter coil in Figure 2 is shown more completely in Figure 3.

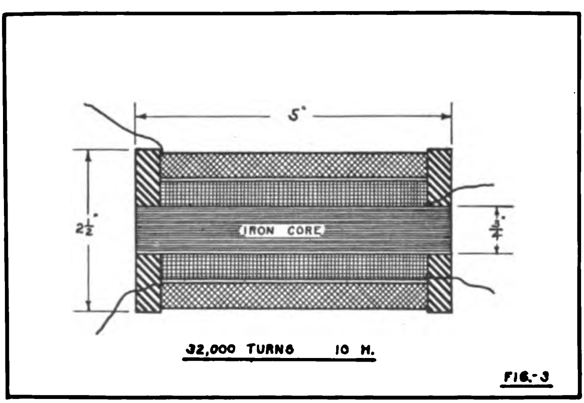


FIGURE 3

[1] De Forest Patent, number 1,201,272, filed September 27, 1915; patented October 17, 1916.

The two windings in each section are the same, the size of the wire being number 34 double cotton covered. A considerable number of filter sets were delivered and, as far as I am aware, they gave excellent results. The coil dimensions were approximately 4 inches long and 2½ inches in diameter. The open core was of soft iron and was 1 inch in diameter.

The first set made was, I believe, delivered to Dr. Goldsmith at the College of the City of New York. One can see even at that time that scientific men foresaw its usefulness, and it was at Dr. Goldsmith's request that he obtained a "B" battery substitute.

As far as I am aware, I never heard of such a circuit being called a filter before that time. I had heard of wave filters in 1914, and presume the name started from that. I recall Eccles' book of December, 1915, showing wave filter circuits. I recognize Campbell's work on wave filters, and believe he has done more to help other inventors than the inventors realize. It is thru the applications of his theories that telephone and telegraph systems have reached the present state of perfection.

Coming back to the circuits that are more than everyday types, I call the attention to Figure 4.

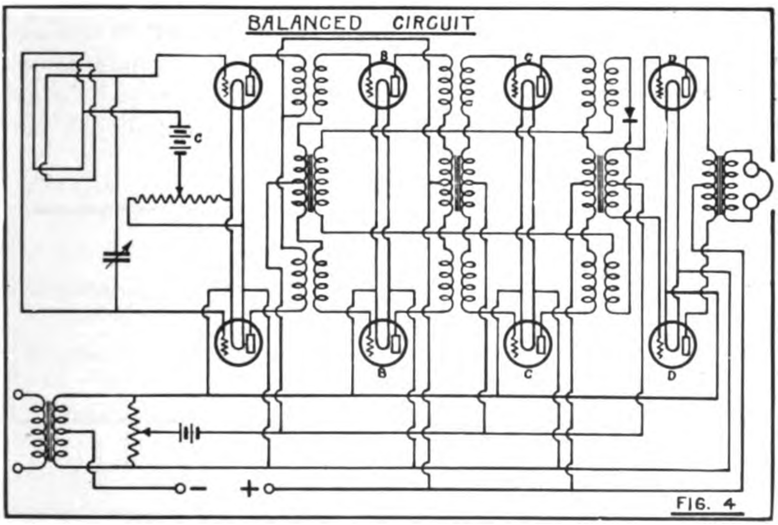


Figure 4

This circuit includes 8 De Forest DV-2 amplifier tubes, one crystal detector, 6 radio frequency transformers, 4 audio frequency transformers, 3 step-up transformers, and one step-down transformer. The step-down transformer was the output trans-

former for the telephones. The three input transformers were balanced type. The "B" battery supply is rectified alternating current, but the filaments were lit with alternating current. Power was delivered from a six-volt step-down transformer.

The general scheme is the "reinforced" or so-called "reflex" method. The balanced circuits are used to oppose changes of induced grid voltage, so that alternating current variations in the filament circuit will not be relayed thru the audio frequency or input transformers. When the tubes are all standard and have similar characteristics this circuit enables good reproduction thru a loud speaker, but at its best the alternating current hum is heard in telephone receivers. Details of this circuit can be obtained on request. Any source of "B" battery can be used in this circuit.

Thermo Junctions

This method of converting alternating to direct current is a most interesting one and should lead to much investigation. It is a simple method yet it is an expensive method of obtaining direct current power. I do not recall any investigation that has held my attention so steadily as this subject. I tried almost all of the common metals one can purchase in quantity in the process of my investigations and I have decided that molybdenum and advance are about as high in efficiency as any other combination. Next came iron and advance, nichrome and advance, and copper and advance. In Figure 5 I show the table in millivolts per junction.

TABLE OF JUNCTION VOLTS
AT RED HEAT

	Molybdenum-advance	Iron-advance	Copper-advance
Milli-volts	100	60	68

Current depends upon cross section of conductor and difference in temperature of hot and cold junctions.
Resistance includes meter resistance and 6 ft. leads.

Figure 5

THERMO ELECTRIC VALUES
See Eccles, 1916, "Wireless Telegraphy and Telephony"

Lead as zero lower limit	18°	Upper Limit	416°
Zinc	373°		
German Silver	175°		

The current thru the hot junction is from the metal of the lower to that of the higher thermo-electric value.

	Microvolts per 1°C.		Microvolts per 1°C.
Iron	$+17.34-.0487\,t$	Silver	$+2.14+.015\,t$
Steel	$+11.39-.0228\,t$	Gold	$+2.83+.0102\,t$
Plat. Ir. alloy 10% Ir.	$+\ 5.96-.0134\,t$	Copper	$+1.36+.0095\,t$
Plat. Soft	$-\ \ .61-.011\,t$	Lead	
Plat. Hard	$+\ \ .26-.0075\,t$	Tin	$-\ .43+.0055\,t$
G. S.	$+12.07-.0512\,t$	Aluminum	$-\ .77+.0039\,t$
Zinc	$+\ 2.34+.024\,t$	Palladium	$-6.25-.0359\,t$
		Nickel to 170°C.	$-22.04-.0512\,t$

(Refer to Eccles' book, cited above, for further data.)

In carrying out the early experiments I had a Weston direct current ammeter with two heavy binding posts. These posts were about 1½ inches apart between centers. I connected the free ends of the junction to each post. Then I would heat the junction until the color was a cherry red. The resistance of the meter was about 0.05 of an ohm and the junction resistance was approximately the same when red hot at the junction end.

I found that the current forced thru by the thermo potential would reach as high a value as 7 amperes for a single junction. I did not take into consideration, however, the cooling of the cold ends by the meter itself until I had made up a battery of junctions containing ten or more. Immediately after I got the hot junctions red hot, the current was no higher than 1 ampere. At first this led me to believe I had reversed junctions; finding this not to be so, I proceeded to cool the cold ends. The more metal cooling surface I added, the greater became the difficulties.

This problem looked possible, but I could not find a way to cool the junctions unless I used a fan or water. This arrangement was then no better than a storage battery.

The next step was to be satisfied with lower potentials and to multiply the number of junctions. The results were just as disappointing as before.

In order to keep the bulk down to reasonable dimensions, I crowded the junctions together and heated them electrically. The more I heated them, the greater became the heat conduction and reflection, thus always keeping the potential lower than I expected.

In Figure 6 I show the first complete 6-volt generator delivering 1 ampere. The power consumed from the line was about

400 watts and I considered it much too high. In this figure is shown the method of getting more heat to the hot junctions than is possible by any other method except that shown in Figure 8

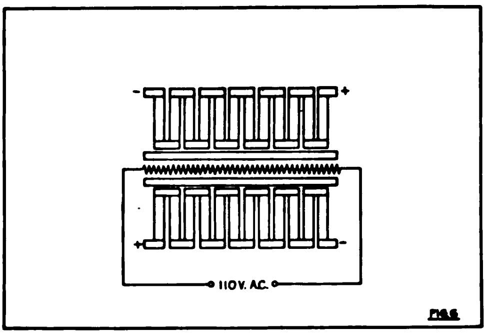


FIGURE 6

A rod of soapstone approximately 6 inches long had been turned down to about 1½ inches diameter with a ¾-inch hole running thru it.

Into this hole I inserted a resistance coil or unit made of nichrome resistance wire, bringing out the ends of the wire at each end of the tube. Two brass plates were fitted on each end and were held together with stay rods separated about ⅜ inch apart. When rows of junctions were laid in between these stay rods and insulated by mica so that no short circuit would occur, each row of junctions were joined together in the proper polarity order. It is obvious that the hot ends of the junctions would rest on the surface of the soapstone. Soapstone has one redeeming feature, and that is, retaining its heat for an almost unbelievable length of time. I believe this method would have succeeded had I been able to find a way of preventing carbonization of the soapstone. If the resistance touched the walls of the soapstone it would invariably short-circuit the turns and burn out.

A very good method of making the junctions for this type of thermo generator is shown in Figure 6A, alternate strips of molybdenum and advance being laid side by side in a holder. Across the ends, from one side to the other, are laid advance strips. In order to weld these securely, the spot welder was used with a hydrogen jet. Whenever the weld was made a blast of hydrogen prevented oxidation. If the hydrogen was not used, an iron flux was used instead. This was a very good method for

welding these widely different strips of metal. When the welds were all made, the sections were cut as shown, the result being a series of junctions.

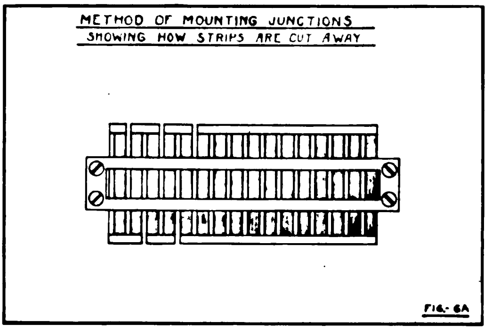


FIGURE 6A

In previous efforts, I had a great many disappointments, because ordinary spot welding would not hold the weld. After this device was constructed and tested, I proceeded in another direction (see Figure 7). This apparatus gave better results than the one in Figure 6 because it was heated by a gas flame and the rising heated air would draw the cool air up with it around the cool ends of the junctions. With a much smaller number of junctions, I could get 8 volts and 1 ampere. This scheme was more economical than the electrical one as it was found to consume only a moderate amount of gas. The more I experimented with this device, the more convinced I became that it is going to be worked out in time.

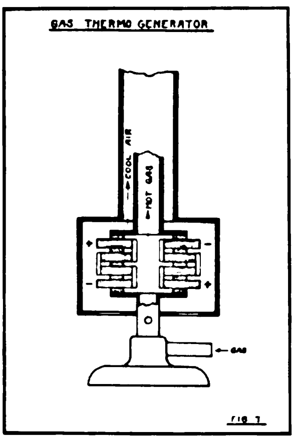


FIGURE 7

Figure 8 shows an induction method which I have worked out, and this, I believe, will eventually lead to the desired results. The drawing in Figure 8 will show some of its general features, but a detailed description is also necessary. Each hot junction is made of copper, and is a complete short-circuited turn. There are 20 of these rings separated by mica insulation. From these rings radiate 16 spokes as in a wheel; 8 are copper and 8 are of another metal such as advance or copel. The copper spokes are welded to the brass ring of its section. The advance spokes connect to the next brass ring. It is obvious, then, that the junction is made at the point where the advance connects to the copper ring. Since the ring is heated by the short circuit, it is directly connected to the junction. The outer ring is split, so that it will not weaken the field unnecessarily.

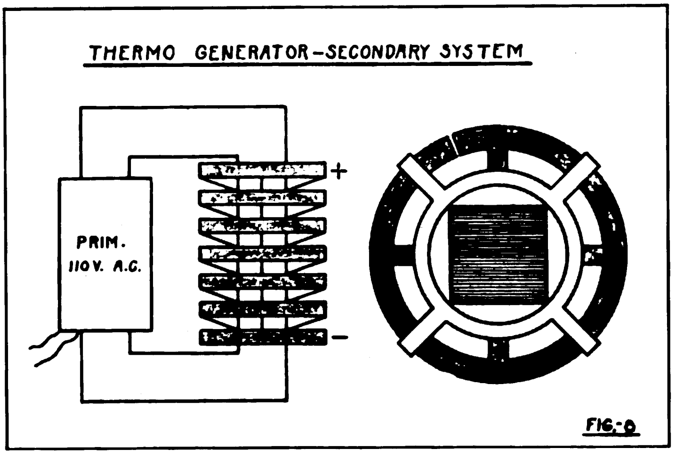


FIGURE 8

RECTIFIERS

Coming to the next step in current supply system I refer again to Figure 2. Here we see one of the early filter circuits. This circuit was the forerunner of Figure 9. Here we have two valves, a condenser, and a transformer, the transformer having two secondaries and one primary winding. It will be noticed that the neutral point of the plate supply is grounded. Connected to this rectifier circuit is a filter system, which embodies the circuit in Figure 2. The constants are noted in the corner of the figure. This is very suitable for operating a detector and two stages of audio frequency amplification.

Figure 10 shows the first of the "B" battery substitutes. It consists of two rectifier tubes, a transformer, and filter systems.

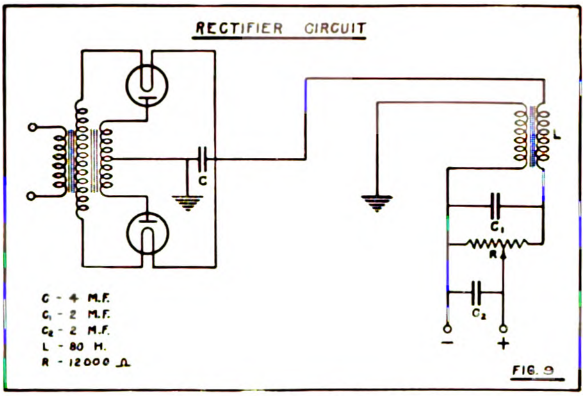


FIGURE 9

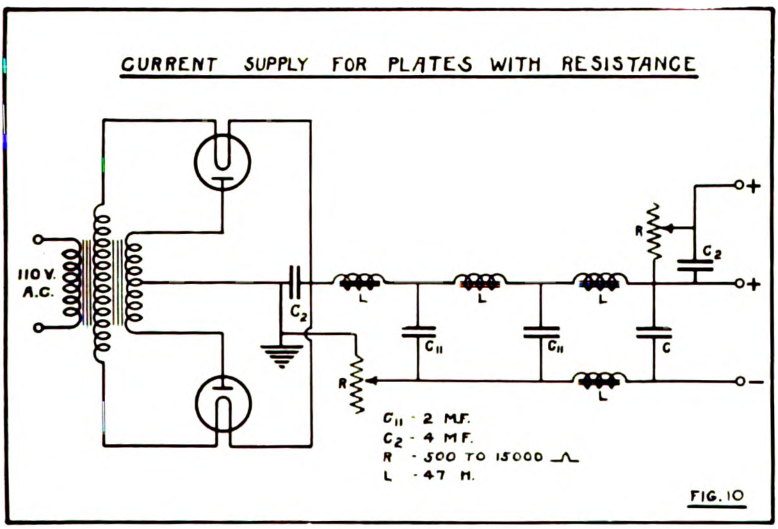


FIGURE 10

It will be noticed that the inductance coils of this circuit are fairly high in value. I found after months of hard work that the line conditions varied considerably, aside from variations in the fundamental frequency itself, and it was not until I combined resistance with capacity and inductance that I was able completely to eliminate all hum even in the most delicate detector adjustments. Many different receiving sets were set up so that comparisons could be made. The simplest filter circuits, such as shown in Figure 10A, were used on circuits that were of the type of neutrodyne, super-heterodyne, straight radio frequency de-

tector and audio frequency stages, but the simple filter circuits would not deliver sufficiently smooth current to a standard De Forest reflex circuit.

Figure 10R is a single-wave rectifier system utilizing two rectifier tubes. It will be noticed that the plate of the second rectifier is in series with the positive side of the rectified current source. This gives a smooth clean current supply entirely free from alternating current hum.

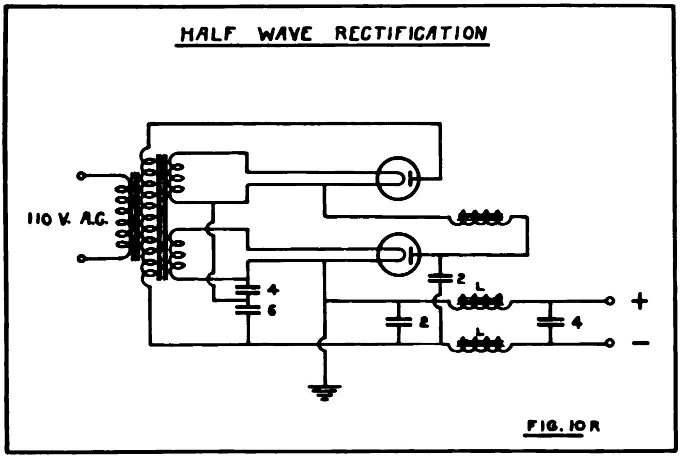


FIGURE 10R

By this time, I had tried the following types of supply on the De Forest reflex circuits:

A. Double filament type.
   This is exceedingly quiet with the audion filaments grounded.

B. Electrolytic single rectification. Best results were obtained by grounding filaments of the audions.

C. Single rectification utilizing tuned filter inductance. Best results with positive line grounded.

This rectifier filter circuit, shown in Figure 10, is efficient and reliable.

The rectifier tube was developed for the circuit, and has proven after months of constant use to be very satisfactory. Each rectifier will successfully pass 40 milliamperes without taking up more than 10 percent of the total emission available. It will be noted that three of the filter coils are in series with the positive leg and one in series with negative line.

It makes little difference where the plate supply circuit is grounded, but for complete elimination of singing I advise

grounding the neutral terminal. If it is desirable to ground the receiver, it is advisable to put a ½ microfarad condenser in series with the receiver ground. A reduced voltage tap is obtained by a high resistance "Bradleystat" (carbon compression rheostat). Shunted across this resistance is a by-pass condenser.

The potentials across the tank condenser C_2 is about 160 volts at a load of 30 milliamperes. The potential across plates and filaments drawing 30 milliamperes is about 125 volts. The no-load voltage across the tank condenser is 260 volts. The terminal alternating voltage is 200 volts. The two rheostats shown in Figure 10 are 10,000 ohm "Bradleystats," consuming 1 to 5 watts.

In Figure 11, I show a more complete rectifier system. One transformer is used. There are two distant rectifier circuits. An auto transformer is made to deliver power to one set of rectifier tubes, and the other coil, a secondary winding, supplies another pair of rectifier tubes. It will be noted that the same

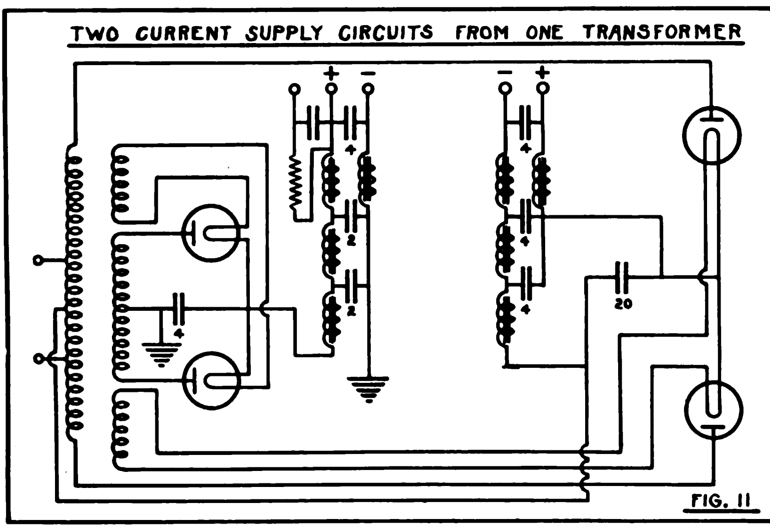


Figure 11

connections are used in Figure 11 as in Figure 10, while the other filter circuit has somewhat similar conditions. This circuit was developed for the purpose of supplying "A" and "B" battery voltage to the special receiving circuit shown in the next figures.

We shall next consider the circuit of Figure 12. This circuit is similar to the one shown in Figure 11 in part, but contains certain other necessary adjuncts. Rectified current is obtained from

two rectifier tubes delivering at full load about 0.3 of an ampere. This current goes to light the filaments of four standard DV-2 tubes which have all been put in series. It is better to connect them in this way as the necessary power is then more easily obtained from the rectifier tubes. There is connected across the terminals of the rectifier circuit a resistance of approximately 3,000 ohms. In order to obtain proper grid bias it is necessary to have a variable contact on this potentiometer. Plate current is obtained either from a rectifier or from "B" batteries. But the primary object is to light the filaments.

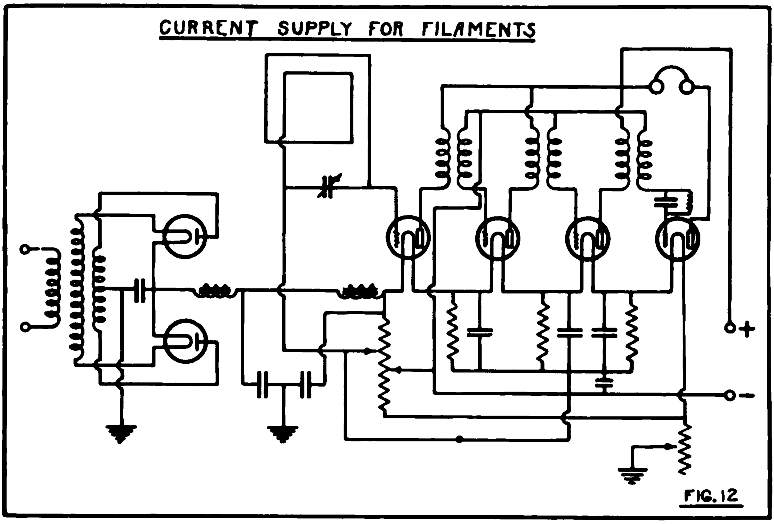


FIGURE 12

It will be noted that starting from one end of each filament is an 800-ohm resistance, all these resistances being connected in common at one end. Across these resistances are 1 or 2 microfarad condensers so that the audio frequency will be distributed equally over the filaments. A condenser connects the grid bias side of the system.

This circuit works very well and gives promise of producing a good system. No disturbances are heard in the receivers. Ground connections are necessary for eliminating ripple.

Radio Frequency Filament Source

There is no doubt that many engineers have tried to utilize radio frequency energy to light the filaments principally because it is so easy to get enough current to light them this way.

Consider the circuit shown in Figure 13. This is not an unsuitable circuit; in fact, it worked better than any other circuit I have tried when using radio frequency energy to light the filaments. However, I found I had a great deal of trouble ahead of me as soon as I tried to tune in for stations. The problem was to find the proper frequency to put thru the filaments. I discovered that the detector was where one fell into most of the trouble. So I shielded the detector and used a separate battery on the filament. Of course, this was not entirely a substitute for batteries, but in these days dry batteries last a long time on one tube consuming only 60 milliamperes. The first frequency used to light the filaments was very high. I found that a frequency of 7,500 kilocycles (40-meter wave length) was useable, but as the frequency was diminished it became more difficult to get results. At first I assumed I would have to make the filament frequency the same as that of the received wave, but that gave too much trouble. There resulted a group of squeals with every twist of the dial.

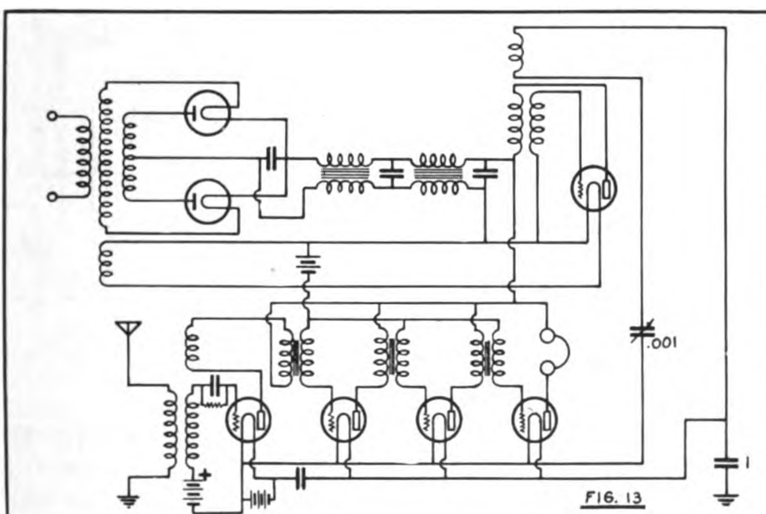


FIGURE 13

Figure 13 will enable one to follow the scheme. I found that the oscillator tube can be a DV-2 as this will supply enough current for 4 DV-3 De Forest tubes. The circuit of Figure 13 is a detector and three stages of audio frequency amplification utilizing regenerative couplings on the detector tube.

Figure 14 is another circuit with which I had considerable success. This scheme was in general a push and pull oscillator;

the rectifier tubes were constructed with grids. The two pancake choke coils in the secondary terminals were to keep the radio frequency out of the transformer. Coupled to the push-pull oscillator is an absorbing circuit which includes the filament circuit. The circuit is tuned to the oscillator frequency, and a separate battery is supplied to light the detector filament. A stopping condenser is inserted to prevent the battery from lighting the filaments of the other tubes. It is very necessary to ground the filaments so that all hum is avoided. A filter circuit is tapped on the tank condenser. All irregular noises are thus ironed out. The generated energy goes to supply the plates of the receiving circuit.

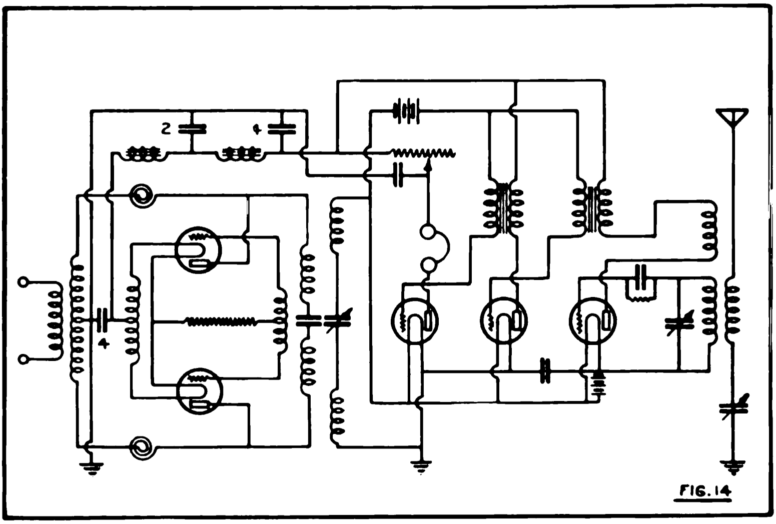


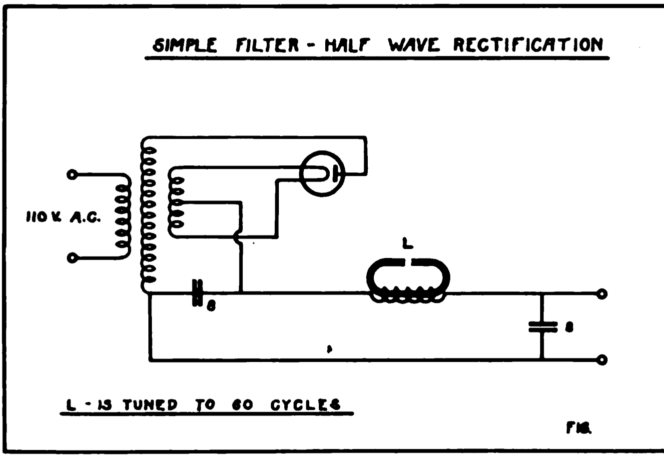
FIGURE 14

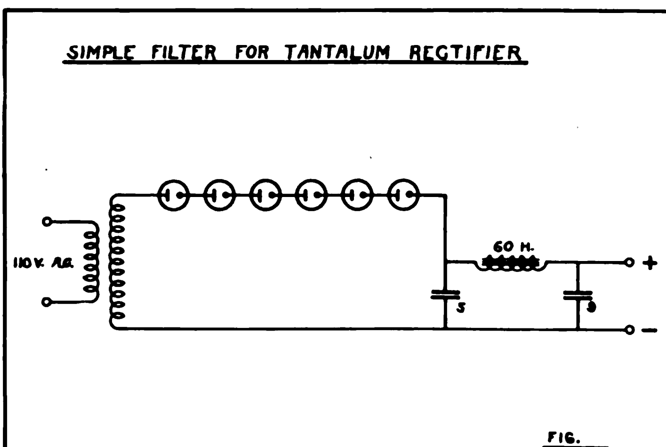
After these tests were completed, I found that in comparison with a standard storage battery, this arrangement was possibly 50 percent as efficient. There is no doubt some disturbing factor which I have made no great effort to locate.

Of course, there is no doubt that there is considerable "paralysis" or blocking, due to effect of the radio frequency voltages acting on the space charge within the tubes. Further than this information, I cannot determine exactly why the arrangement works at all since we know well that such rapid reversals set up opposing forces within the tube.

The detector was my stumbling block. The amplifier was prejudicially affected by the radio frequency currents used, be-

cause when I substituted a storage battery supply, the amplification was noticeably increased.





I shall now describe an entirely different idea in current supply devices, and one which is indeed very attractive. It will deliver both "A" and "B" voltages constantly. The efficiency is fair. It consumes about 200 watts from the line and will deliver 50 to the receiving circuit. Lengthy tests have not been carried on to any great extent, but even from the time it was first tested it convinced our engineers that it had great merits. The plate voltage is fairly constant, and the device will deliver 150 volts and up to 50 milliamperes with real reliability. In tests

made, I have found that there is no ripple in the plate circuit, but the filament supply gave us some concern until the proper filter was found.

While at this time we cannot tell much about its life, we presume it will turn out to be satisfactory in this respect. Owing to the continued development of this outfit by the manufacturer, we are prevented from giving full details.

The outfit is a miniature motor-generator set. Its speed is 6,000 revolutions per minute. There are 24 bars on the commutators. The fields are excited by the high voltage side, and the motor is universal for 110 volt alternating or direct current. The commutator ripple has a frequency of 2,400 cycles per second, and this frequency is easily filtered out with proper condensers and impedances. The bearings are sleeve wick lubricated, which probably is better than ball bearings. They need but little attention when once the cups are filled with oil. The motor is cooled by a fan attached on the shaft. The plate voltage across the commutator is 150 volts. The filament supply is 8 volts. No commutator sparking disturbance, due to the power line, is heard in the receiver. The little device is also a fine outfit for charging storage batteries as it has the two separate voltages.

I should like to see a perfected "A" battery substitute supply that will light the filaments as well as a motor generator can be made to do, and without having to change the accepted multiple connection of filaments. The filter on the low voltage side would be a very big affair if we tried to make it do as well as the "B" battery substitute. If we found we needed a 20-henry choke coil and suitable condenser to smooth out the ripple from a Tungar rectifier, we would have to use a bulky outfit. I do not know whether there are any electrolytic "A" battery substitutes on the market at this time, but I believe the problem is just as great as when rectifier tubes are employed to rectify the alternating current supply. It would be interesting to hear from anyone concerning secondary emission tubes, such as Dr. Hull proposed some time back.

SUMMARY: There are considered various forms of "A" and "B" battery substitutes such as rectified and filtered alternating current generators, thermo electric devices, radio frequency current generators, electrolytic rectifiers, and motor generators.

A METHOD OF MEASURING RADIO FREQUENCY BY MEANS OF A HARMONIC GENERATOR*

By

August Hund

(Radio Laboratory, Bureau of Standards)

Introduction

Harmonics of an electron tube generator have been used for radio-frequency measurements[1] for several years. The well-known system for measuring a radio frequency in terms of an audio frequency, devised by Abraham and Bloch[2] and employing their multi-vibrator is perhaps the best example of such use of harmonics.

It is the purpose of this paper to describe briefly an arrangement which employs a simple type of harmonic generator the fundamental of which is an audio-frequency alternating current. The generator is rich in harmonics of sufficient power to produce appreciable currents in a wavemeter which is coupled to it and tuned to resonance with frequencies which are integral multiples of the fundamental. The arrangement is a convenient one for primary frequency standardization since the fundamental frequency produced by the harmonic generator can be checked during the measurement against the frequency of a standard tuning fork by means of a visually indicating instrument. A large number of harmonics can be utilized since decidedly sharp settings are made possible by the use of another visual indicator. The visual indicators lighten the labor of the observer since it is found easier to take a large number of settings with the eye than by ear.

Description of the Method

The arrangement for primary frequency standardization is shown in Figure 1. The heavy line portions are the essential parts

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[1] "Radio Instruments and Measurements," Bureau of Standards Circular Number 74, pages 100–104 ; 1918.

[2] "Comptes Rendus," volume 168, page 1105 ; June, 1919.

of the system. The visual beat indicator is a portable galvanometer, with 1 milliampere for full scale deflection. It is used for keeping the fundamental frequency f of the harmonic generator exactly at the value of the frequency of the standard tuning fork generator. The pointer will stand still when the two frequencies are exactly alike. This will also happen when the

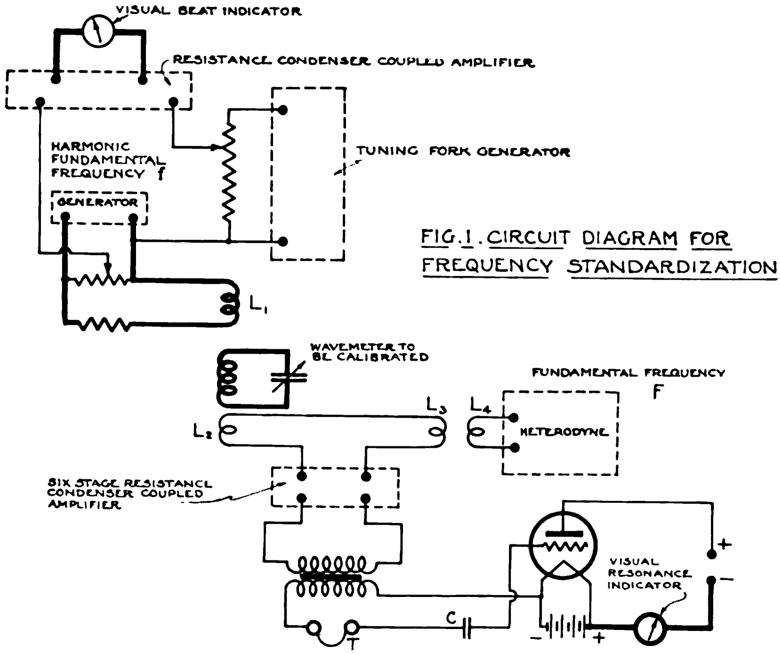

FIGURE 1

two frequencies are very different, but such a shift of frequency is not likely to happen during the measurement and after having the fundamental frequency f adjusted. The visual indicator of resonance will indicate a sharp minimum whenever the wavemeter is tuned to the fundamental frequency f or any integral multiple of it. The procedure of measurement is briefly as follows :

(1) The fundamental frequency f of the harmonic generator is varied until the fundamental alternating current is in synchronism with the current produced by the standard fork generator. The zero beat indicator will then stand still and the least change in the frequency will cause the pointer to swing[3]

[3] Two swings per second, for instance, mean that the fundamental is off by two cycles. The pointer will, of course, also stand still when the harmonic generator is adjusted to $2f$, $3f$, and so on.

to and fro. The frequencies of all possible currents have then the form ($a \times f$) if a denotes whole numbers such as 1, 2, 3, 4, and so on.

(2) The heterodyne is set to certain fundamental frequency F which is in the neighborhood of the frequency range to be used in the calibration.

(3) While the wavemeter is varied, a series of beat notes will be heard in the telephone receiver T which produce at the same time minimum deflections on the resonance indicator. The telephone receiver is used as a rough guide and the settings are made by means of the visible indications. The minimum deflections will occur whenever

$$a \cdot f = b \cdot F$$

if $b.F$ stands for all possible frequencies of the heterodyne and b denotes integer numbers such as 1, 2, 3, 4 . . . . . This shows that a minimum indication is obtained whenever the wavemeter is set to a frequency which is an integral multiple a of the fundamental frequency f of the harmonic generator. Beats can also occur between harmonics of the harmonic generator and harmonics of the heterodyne, which is expressed by factors a and b. The factor b can be made equal to unity by choosing a rather loose coupling between the heterodyne and the six-stage amplifier. Such a precaution is, however, not necessary.

(4) Knowing approximately the fundamental frequency F of the heterodyne driver it is easy to determine the particular harmonic frequency $a \times f$ to which the wavemeter is tuned. The simplest way, however, is to substitute the wavemeter by one the calibration of which is known approximately. (Example : Suppose the fundamental frequency $f = 1$kc. (kilocycle per second) and the approximate frequency is found to be $a.f = 21.13$kc., then the true setting is 21 kc., since a must be a whole number).

When the precautions above mentioned are followed, this method may be used to obtain resonance settings on a wavemeter up to the 100th harmonic. It has been found possible to obtain as high as the 360th harmonic of a 1,000-cycle fundamental frequency, but this requires much skill and does not appear useful in actual wavemeter standardization work. For the higher frequencies it is preferable to use a tuning fork of higher fundamental frequency f, or to employ an additional harmonic generator and measure its fundamental frequency f' against a certain harmonic ($a \times f$) by another zero beat indicator in the same way as f is checked against the fork driver. An an example, suppose the fundamental frequency f of the first harmonic

generator is adjusted to the frequency of the tuning fork, which let us assume to be $f = 1$ kc. The fundamental frequency f' of the second harmonic generator may be adjusted by another zero beat indicator to the 20th harmonic of f, that is, $f' = 20$ kc. A calibration of a wavemeter at 500 kc. would then require only the 25th harmonic of the second harmonic generator instead of the 500th harmonic of f.

Practical Hints and Description of Apparatus

(I) The *harmonic generator* is shown in Figure 2. It makes use of the fact that a circuit of this type produces strong harmonics and especially when much inductance with pronounced distributed capacity is used in the grid and plate circuits and a crystal rectifier in the output branch. In order to accomplish this a hard rubber tube, one-half inch in diameter and five inches in length was used as a core for a coil with 48,000 turns of wire. Using wire, about Number 30, American wire gauge, and winding the coil in the ordinary way will give pronounced coil capacity. Taps were taken off every 3,000 turns as indicated in the figure, giving enough sections of different inductance to produce distorted audio currents ranging from about 10 cycles per second to 15 kilocycles per second. An ordinary air condenser connected from the plate to the grid gives a means for varying the frequency gradually within a small range. A tube capable of giving from 2 to 10 watts power is sufficient for the measurement. Two tubes connected in parallel can also be used. The entire generator should be shielded by copper foil with the output coil outside of the screen.

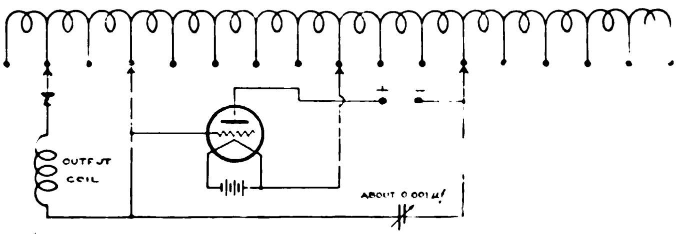

FIGURE 2—Diagram of Connections for the Harmonic Generator

(II) The 6-*stage amplifier* (Figure 3) is used for amplifying the beat effects between the current of frequency *a.f* to which the wavemeter is tuned and the current produced by the heterodyne. It acts normally as an audio-frequency repeater since the beat notes are usually within the audible range of frequencies. It operates as a radio-frequency amplifier when no heterodyne

is used. This is only possible for comparatively low frequencies (not much higher than the 20th harmonic of f). It seems therefore best to use the heterodyne for all settings. The entire arrangement is surrounded by a shielding of copper foil. If possible the " B " battery should be within the shielding, otherwise a by-pass condenser should be used within the screen and across the terminals leading to the " B " battery.

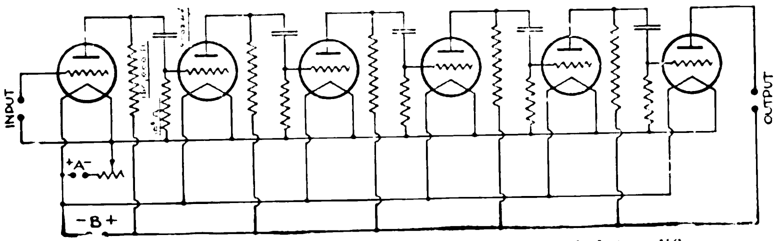

FIGURE 3—Six-Stage Resistance Condenser Coupled Amplifier

(III) The *amplifier* leading to the visual beat indicator is likewise resistance-condenser coupled and shown in Figure 4.

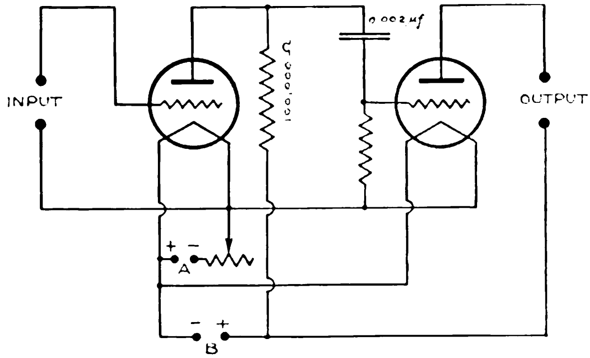

FIGURE 4—Resistance Coupled Amplifier for Zero Beat Indicator

(IV) The circuits leading to the *visual resonance indicator* are shown in Figure 1. An ordinary audio-frequency transformer is employed for coupling the output branch of the 6-stage amplifier to the circuit with the visual resonance indicator. It was found that the resonance settings can be made sharper when a dip method is used. Such an indication is obtained when a grid condenser of about 0.0001 μf. is employed without a leak in the detector circuit. When the different beat notes affect the detector tube the microammeter in the plate circuit will show a sharp dip. If the frequency measurement is carried out in a

laboratory where many outside disturbances affect the grid, the telephone receiver T (Figure 1) indicates the presence of interfering voltages. Under such circumstances the use of a grid leak is advantageous.

(V) The *tuning fork generator* provides the fundamental frequency in terms of which the measurement is made. It is therefore essential to use a standard fork the frequency of which is constant and known. It seems best to use a fork employing an electron tube drive of such a type that the circuits can produce alternating current only when the fork is vibrating at its own natural frequency. For most work a fork having a frequency of about 1,000 vibrations per second is convenient. A fork giving about 100 vibrations per second would have the advantage of giving more points on the wavemeter to be calibrated, but the disadvantage of not giving calibrations at frequencies as high as can be obtained with a 1,000-cycle fork.

(VI) The *heterodyne* is an ordinary electron tube generator covering a range of frequencies from about 10 to 1,500 kc. Copper foil is used for shielding the entire apparatus except the coil L_4 (Figure 1) which couples to the 6-stage amplifier. A tube giving about one-tenth of a watt radio frequency power will do.

(VII) The inductances of the *coupling coils* L_1, L_2, L_3 and L_4 shown in Figure 1 depend on the range of frequency used, while the shape of coils L_1 and L_2 depends, besides, on the shape of the coil used in the wavemeter. A loose coupling to the wavemeter is essential. The best distance between L_1 and the wavemeter, L_2 and the wavemeter, and between L_3 and L_4 is found by trial for which case it is convenient to use the telephone receiver T as well as the resonance indicator because the relative positions of the coils may be so much off that the indicator will not respond at all. The coils L_3 and L_4 are made to slide in a box which is covered with copper foil.

(VIII) The *zero beat indicator* in Figure 1 can also be replaced by a cathode-ray oscillograph (the hot-cathode type is convenient). The usual way of using this apparatus applies the two voltages E_1 and E_2 coming from the harmonic generator and fork generator, respectively, to the two deflection condensers of the oscillograph. Synchronism is then recognized by a stationery pattern (Lissajous figure) on the fluorescent screen of the cathode-ray tube. Another way is partly indicated in Figure 5. The current coming from the fork generator produces a circle on the screen of the cathode-ray tube. The battery between the hot cathode and the anode of the oscillograph is connected

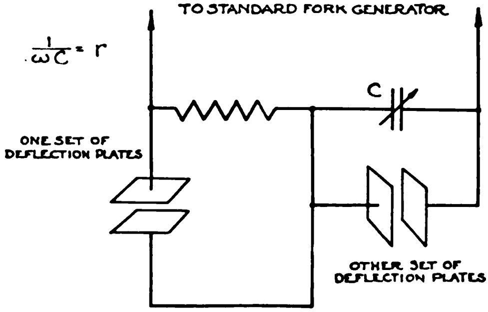

FIGURE 5—Zero Beat Indicator, Using a Cathode Ray Tube

in series with the output of the harmonic generator. If the fundamental frequency of this generator is not exactly the same as the frequency of the fork, the diameter will increase and decrease at the rate of the frqeuency difference.

Department of Commerce,
Washington, D.C.
September 2, 1924

SUMMARY: An improved method has been developed for standardizing a wavemeter by means of the harmonics produced by a simple type of harmonic generator the fundamental of which is an audio-frequency alternating current. The arrangement is adapted to primary frequency standardization since the fundamental frequency can be checked during the measurement against the frequency of a standard tuning fork by means of an instrument using visual indication. A large number of harmonics (up to 100 and higher) can be utilized since very sharp settings are made possible by the use of a visual resonance indicator.

AN ELECTROMETER METHOD FOR THE MEASUREMENT OF RADIO FREQUENCY RESISTANCE*

By

P. O. Pedersen

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Introduction

To determine the resistance of a conductor for direct current is one of the simplest electrical measurements and can be carried out with a very high degree of accuracy. The case is entirely different when measuring the effective resistance for radio frequency current.

To determine this resistance is still a rather difficult matter and cannot be done with any high degree of accuracy, altho the development of suitable generators for the production of continuous radio frequency current has resulted in considerable improvement.

The older methods[1] practically all rested on the fundamental investigations of *V. Bjerknes* on resonance in simple and coupled circuits[2]. Measurements were generally carried out in the way that the circuit to be investigated was set in oscillation by means

*Received by the Editor, October 16, 1924.

[1] E. Nesper: "Die Frequenzmesser und Dämpfungsmesser der Strahlentelegraphie," pages 165-238. (Leipzig, 1907.)

[2] V. Bjerknes: (a) "Dämpfung schneller electrischer Schwingungen," 'Wied. Ann," 44, page, 14, 1891.

(b) "Ueber den zeitlichen Verlauf der Schwingungen im primären Hertzschen Leiter," "Wied. Ann.," 44, page 513, 1891.

(c) "Ueber electrische Resonanz," "Wied. Ann.," 55, page 121, 1895.

of a spark-discharge, and the oscillations then investigated in an auxiliary circuit loosely coupled to and in resonance with the circuit under test. As indicator in the auxiliary circuit there was generally used either a quadrant electrometer the deflection of which is proportional to the time integral of the square of the potential difference or, and in most cases, a radio frequency ammeter the deflection of which is proportional to the square of the current.

Most of the newer methods[3] are based upon the use of continuous oscillations which originate in a generator circuit and by means of suitable coupling arrangements act upon the test circuit—namely, the circuit under test—and in this case the measurements are carried out directly in the test circuit.

Figures 1 and 2 show the principles of the two methods now nearly always used. In both figures, *G H F* stand for generators of continuous radio frequency current, while *S* indicates coupling arrangements, *C* variable air-condensers, *A* ammeters for radio frequency currents, and *r* variable resistances of which the values of the various settings are accurately known.

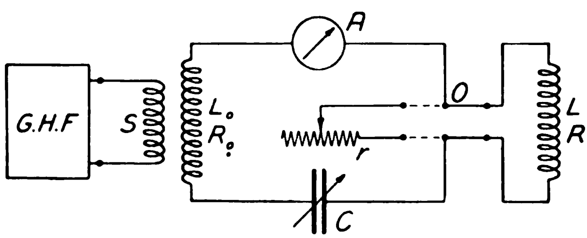


FIGURE 1—The Substitution Method

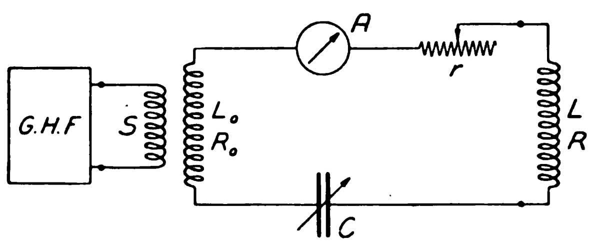


FIGURE 2—The Added-Resistance Method

[3] (a) J. H. Dellinger: "The Measurement of Radio Frequency Resistance," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, New York, volume 7, pages 27-60, 1919.

(b) "Radio Instruments and Measurements," "Bureau of Standards Circular," Number 74, Washington, 1920.

(c) H. Armagnat et L. Brillouin: "Les mesures en haute frequençe," Ecole Supérieure d'Electricité, Section de Radiotélégraphie," Paris, 1924.

Figure 1 shows the substitution method. The effective radio frequency resistance of the wire coil L is measured in the following manner: With the switch O in the position shown, the condenser C is varied until the circuit $C\ L_o\ A\ O\ L$ is in resonance with the generator-circuit, which is the case when the reading, A_{max}, of the ammeter is a maximum. Next the switch O is changed over to the position shown in broken lines and then the circuit $C\,L_o\,A\,O\,r$ is brought into resonance with the generator circuit. The reading A_{max}' of the ammeter will in this case generally be different from the reading A_{max} first determined; but by varying the value of r we may arrange that $A_{max}' = A_{max}$. In that case we have $R = r_o$ where r_o is that value of r for which the reading is the same in both cases.

Figure 2 shows the added-resistance method. For a given value r' of r the circuit $C\,L_o\,A\,r\,L$ is brought into resonance, just as described above, and the corresponding reading A_{max}' is taken. This is repeated for a value r'' with a corresponding reading A_{max}''. We then have the total resistance of the circuit $R^o = R_o + R$ determined by

$$R^o = \frac{r''\,A_{max}'' - r'\,A_{max}'}{A_{max}' - A_{max}''}. \tag{1}$$

In carrying out these measurements, the following difficulties and sources of errors are, among others, encountered:

1. The emf. must not vary during a test. This requirement is, however, often difficult to fulfil, especially when using the method of added resistance where the reaction of the test circuit on the generator circuit varies with the different settings of the resistance.

2. The coupling should be loose and must not vary during a test. This is often difficult to attain, especially with the substitution method where the coupling between the generator circuit and the auxiliary circuit often will be partly of electrostatic nature, and this part of the coupling may vary in an uncontrollable manner when throwing over from the coil under measurement $(L\ R)$ to the known resistance (r).

3. The readings are quite difficult to make and take a comparatively long time.

For circuits having some $0.3 - 1.0$ ohms, one must be very careful in order to obtain a mean-error not over 2 percent, and in many cases the errors surely reaches 10 percent or even considerably more.

In several respects it would be an advantage to have to deal

only with the test circuit itself and its oscillations, as in that case no difficulties would arise from the coupling between the test and the auxiliary circuit or the generator circuit.

For oscillations in a single circuit containing the condenser C, the inductance L, and the resistance R (see Figure 4, heavy line circuit), the current i is given by

$$\left.\begin{aligned} i &= \frac{h^2}{\kappa} V_o C \, \varepsilon^{-\kappa t} \sin \nu t = \\ &\frac{V_o}{\sqrt{\frac{L}{C} - \frac{R^2}{4}}} \cdot \varepsilon^{-\frac{R}{2L}t} \sin\left(t\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}\right), \end{aligned}\right\} \quad (2)$$

where $\kappa = \frac{R}{2L}$, $h^2 = \frac{1}{LC}$, $\nu = \sqrt{h^2 - \kappa^2}$ and where V_o is the potential of the condenser at the beginning of the oscillatory discharge. Figure 3 represents a current curve obtained according to formula (2).

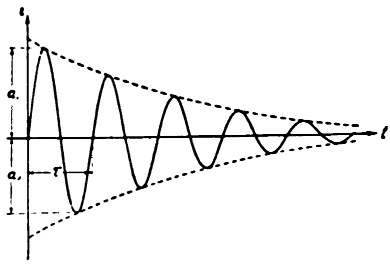


FIGURE 3—Curve of Oscillations

For feebly damped oscillations, formula (2) may, to a close approximation, be written:

$$i = V_o \sqrt{\frac{C}{L}} \cdot \varepsilon^{-\kappa t} \sin ht = V_o \sqrt{\frac{C}{L}} \cdot \varepsilon^{\frac{R}{2L}t} \sin \frac{t}{\sqrt{CL}}. \quad (2, 1)$$

If the effective resistance R of the circuit is to be determined only by means of the oscillations in the circuit itself, without any auxiliary circuit, this must be done by utilizing in some way or other the natural constants of the oscillations. There are here several possibilities. If, for example, $\frac{a_1}{a_2}$ is the ratio between the maxima of two successive amplitudes in opposite directions (see Figure 3), then we have

$$\frac{a_1}{a_2} = \varepsilon^{\frac{\pi}{2} R \sqrt{\frac{C}{L}}},$$

or

$$R = \frac{2}{\pi} \sqrt{\frac{L}{C}} \log_e \frac{a_1}{a_2} \quad (3)$$

Knowing the value of L, C and the ratio $\frac{a_1}{a_2}$, R may be calculated.

This method was used by E. Rutherford[4] and J. Zenneck.[5] By the former, the ratio $\frac{a_1}{a_2}$ is determined in a very ingenious manner by means of the magnetic influence of radio frequency current on magnetically saturated steel needles, while the latter uses a Braun-tube for the determination of the ratio between the successive maximum amplitudes of the condenser potential.

Both of these methods are, however, rather inconvenient and inaccurate. Neither of them is suitable for feebly damped circuits.

There is another way, however, to utilize the constants of the current curve for the determination of R, which leads to a method having considerable advantages as compared with the methods so far used.

This method has been developed in the Laboratory of Telegraphy and Telephony of the Royal Technical College, Copenhagen, and the principle underlying it has previously been published elsewhere.[6]

1. Theory of the New Method

We will consider the arrangement shown in Figure 4. The condenser C, assumed for the present to be without leakage, is charged to the potential V_o when the key N is in the position ac. If now the key is changed over to the position ab, then the charge of electricity CV_o will be discharged thru the circuit shown in heavy lines and the value of the current as a function of time is given by formula (2).

The potential difference between A and B is then $L\frac{di}{dt}$, and the time-integral D_0 of the square of this potential from $t=0$ to $t=\infty$ is then

[4] E. Rutherford: "A Magnetic Detector of Electrical Waves and Some of Its Applications," "Phil. Trans.," A, volume 189, page 1, 1897.

[5] J. Zenneck: "Verfahren, um die Dämpfung elektrischer Schwingungen sichtbar zu machen," "Ann. d. Phys.," Volume 7, page 801, 1902.

[6] P. O. Pedersen: (a) "Metode til Bestemmelse af den effektive Modstand i højfrekvente Svingningskredse," "Vid. Selsk. Math.-fys. Medd.," IV, 5, 1922, Copenhagen.

(b) "A Method for the Measurement of Radio Frequency Resistances," "Wireless World and Radio Review," page 135, April 29, 1922.

(c) "En ny Metode til Bestemmelse af den effektive Modstand i Højfrekvente Svingningskredse," "Ingeniøren," pages 185-196, April 19, 1924.

$$D_o = \int_o^\infty \left(L\frac{d\,i}{d\,t}\right)^2 \cdot dt = \frac{V_o^2}{4\kappa} = V_o^2 \frac{L}{2R}, \tag{4}$$

or

$$R = \frac{L}{2\,D_o} V_o^2 = \frac{L}{2\beta P_o} V_o^2 \tag{4, 1}$$

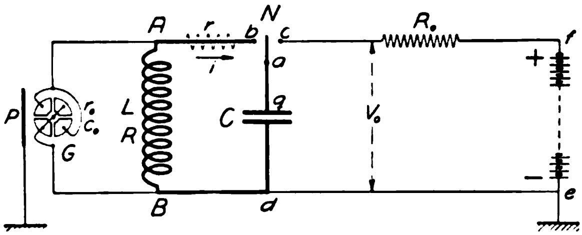


FIGURE 4—Schematic Representation of the Arrangement. (The test-circuit is shown in heavy lines, G is a Quadrant-Electrometer connected up as shown and resting on a grounded plate P (see also Figure 5). The one terminal e of the battery is also grounded)

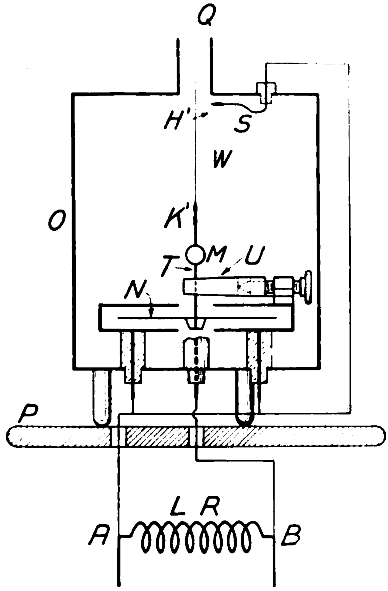


FIGURE 5a—The Quadrant Electrometer G. The needle N carried by the aluminum wire T with the attached mirror M, is suspended by means of the quartz fiber Q. Connection is effected thru the stout wire S and the thin Wollaston wire W—S being insulated from the cover by an amber bushing. The cover is connected to earth over the plate P. U is a clamping arrangement for the needle

The quantity D_0 may be measured, for example, by means of a quadrant electrometer G inserted between A and B in such a manner that one pair of oppositely placed quadrants and the needle are connected to one of these points while the other pair of quadrants is connected to the other point as shown in Figures 4 and 5.

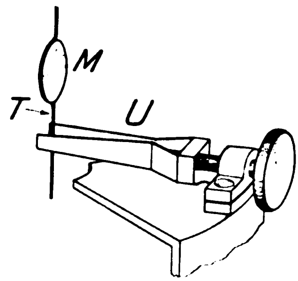


FIGURE 5b—Details of clamping arrangement. The tweezer-form spring U is made of phosphor-bronze and is opened and closed by means of a conical-pointed screw in such a way that on closing, the wire T with needle and mirror is raised a little by the spring

The throw P will then be proportional to D. The determination of the ballistic constant β will be discussed later on.

If the values V_o, L, β, and P_o are known, equation (4, 1) gives the value of R.

If V_o and L are kept constant and an extra resistance r is inserted in the circuit (see Figure 4) we have—if P_r is the value of the throw corresponding to the resistance $R+r$,—

$$R+r=\frac{L}{2\beta P_r}\cdot V_o^2, \tag{4, 2}$$

and from this and equation (4, 1)

$$\frac{R+r}{R}=\frac{P_o}{P_r} \text{ or } R=r\frac{P_r}{P_o-P_r} \tag{5}$$

By means of two measurements—one with and one without the added resistance r in the circuit—R may be determined without knowing the values of V_o, L or the ballistic constant β of the galvanometer.

2. INVESTIGATION OF THE VARIOUS POSSIBLE SOURCES OF ERROR

If the condenser has any leakage, it will lose some of its charge in the time interval from the moment N breaks its con-

nection at c till it makes connection at b (see Figure 4). If the leakage of the condenser is considerable and if the said time interval is not very short, a considerable error may be caused thereby. As will be shown in the following, various difficulties are encountered in constructing the key N in such a way that the above-mentioned time interval will be very short. The difficulty resulting from this, may, however, be entirely eliminated by using the test arrangement shown in Figure 6.

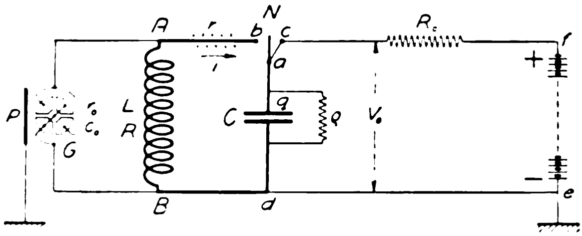


FIGURE 6—Schematic View of Test Arrangement for Oscillating Circuits Containing a Condenser Having a Leakage, ϱ ohms

Here the terminals a and c are connected together so that the terminals of the condenser C having a leakage of ϱ ohms are permanently connected to the terminals e and f of the battery. The discharge is then started immediately when the contact ab is made, and at the beginning of the discharge the condenser has the charge $V_o C$—assuming that $\varrho \gg R_o$, where R_o is a large resistance inserted in the lead cf.[7]

The logarithic decrement δ of an oscillating circuit having a series resistance R and a condenser shunted with ϱ ohms is known to be:

$$\delta = \pi \sqrt{\frac{C}{L}} \left(R + \frac{1}{\varrho} \cdot \frac{L}{C} \right). \tag{6}$$

By connecting the terminals of the battery permanently to the condenser the decrement is accordingly increased to:

$$\delta' = \pi \sqrt{\frac{C}{L}} \left(R + \frac{1}{\varrho} \cdot \frac{L}{C} + \frac{1}{R_o} \cdot \frac{L}{C} \right), \tag{7}$$

and consequently the effective resistance of the circuit, owing to the permanent connection between condenser and battery, is increased by the amount

$$r^{\prime\prime} = \frac{1}{R_o} \cdot \frac{L}{C} \tag{8}$$

[7] If a very large leakage-free condenser is at hand (see part 4), the arrangement of Figure 4 may be used even for a leaky condenser C, provided that the leakage-free condenser is inserted in series with the leaky condenser.

By the measurement under discussion, there is directly determined the total resistance

$$R^o = R + \frac{1}{\varrho} \cdot \frac{L}{C} + \frac{1}{R_o} \cdot \frac{L}{C}. \tag{9}$$

In order to derive the actual equivalent resistance

$$R' = R + \frac{1}{\varrho} \cdot \frac{L}{C} \tag{10}$$

of the circuit itself, the correction r^o determined by means of formula (8) should be deducted from the resistance R^o determined by the throw of the galvanometer.

In order to keep this correction small, it is advisable to give R_o a rather high value. On the other hand it is necessary that $R_0 \ll \varrho$ in order that the condenser potential may, to a close approximation, be put equal to V_o. (The resistance R_o should be inserted in the lead $c\,f$, and not in $d\,e$. In the latter case, the zero position of the electrometer would be displaced during a measurement, since the potential of the two quadrants and the needle before the discharge would be very nearly zero, but during the discharge would be almost equal to V_o).

The correction calculated according to (8) would, for two reasons, be a little too large. Firstly, the resistance R_o will always have a little inductance, and consequently its impedance is larger than R_o. Secondly, R_o will not be perfectly free from capacity and this capacity added to the capacity between the leads $c\,f$ and $d\,e$ (see Figure 6) will increase the capacity of the oscillating circuit. As the frequency of the oscillations is thereby reduced, the effective resistance of the circuit will also be reduced a little. In general, however, both of these corrections will be exceedingly small.

In the measurements on leaky condensers described in the earliest publications (6, *a* and *b*) large choke coils and large resistances were inserted in both of the battery leads, and therefore no correction was made for the additional damping caused by the battery leads. There is no doubt, however, that it is more rational—as was done by Nancarrow and Cohen[8]—to dispense with the choke coils and to correct for the increase in damping. For the reason stated above, it is better, however, to insert the resistance in one of the battery leads only, connecting the free end of the battery with the key—and not in both battery-leads as shown in Figure 2 of the paper of Nancarrow and Cohen.

[8] F. E. Nancarrow and I. J. Cohen, "High Frequency Resistance Measurement," "Post Office El. Eng. Journal," volume 16, pages 71-81, April, 1923.

In the case shown in Figure 6, an additional current will flow thru the inductance L as long as $a\ b$ is closed. The value of this current is V_o/R_o, and if we call the ohmic resistance of L, r', then the result will be a potential difference $V_o \frac{r'}{R_o}$ between the points A and B. This potential difference will, in normal manner, act upon the quadrant electrometer, but is so small that its influence is negligible. If, for instance, $V_o = 500$ volts, $R_0 = 50{,}000$ ohms, $r' = 1$ ohm, and if the key is closed 10 seconds, then the corresponding value of $(\text{potential})^2 \times \text{time} = 1 \times 10^{-3}$ volts^2 seconds. The value corresponding to the oscillations is generally more than 20 volts^2 × seconds. The error caused by the direct current thru L is therefore quite unimportant. The resistance r should, of course, not be inserted between the two points A and B to which the quadrant-electrometer is connected.

The preceding considerations and corrections apply only to the measurement of condensers having so great a leakage that the arrangement shown in Figure 6 must be applied.

A further number of possible sources of error found when applying this method of measurements must be considered—and this even in those cases where the more simple arrangement shown in Figure 4 may be used.

In the derivation of formula (4), only the inductance L has been considered, but not the effective resistance of the coil. It is very easily seen, however, that the error arising from this—for all circuits which are not extremely highly damped—is of no importance.

Furthermore there is the possibility that the appreciable resistance r_o of the lead to the electrometer needle in connection with the capacity c_o between the needle and the two quadrants and the opposing pair of quadrants, respectively, may cause a not insignificant error, partly by increasing the damping of the circuit and partly by reducing the potential of the needle.

The first-mentioned error causes an increase R' of the effective resistance, which, to a close approximation, may be determined by:

$$R' = r_o \frac{c_o^2}{C^2}, \tag{11}$$

as c_o is so small that practically the entire voltage drop is across c_o. This is easily seen from the following: If the potential v between A and B is given by (see Figure 4)

$$v = V_m \sin \omega t, \tag{11, 1}$$

then the current thru the condenser C and the inductance L is determined by

$$i = \omega V_m C \cos \omega t \tag{11, 2}$$

while i' thru the electrometer lead under these conditions is given closely enough by

$$i' = \omega V_m c_o \cos \omega t \tag{11, 3}$$

The loss of power in the oscillating circuit is consequently equal to $\frac{1}{2} \omega^2 V_m^2 C^2 R$ and in the electrometer $\frac{1}{2} \omega^2 V_m^2 C_o^2 r_o$. The total loss is therefore

$$\tfrac{1}{2} \omega^2 V_m^2 (C^2 R + c_o^2 r_o) = \tfrac{1}{2} \omega^2 V_m^2 C^2 (R + R')$$

where R' has the value determined by (11). The capacity of the needle does not necessarily exceed 5 cm. and the resistance r_o of the needle-lead may be about 200 ohms, while the condenser C will often be over 10,000 cm. If $c_o = 5$ cm., $r_0 = 200$ ohms, and $C = 10{,}000$ cm., then $R' = 0.00005$ ohms—a quite negligible increase in the effective resistance of the circuit.

The fraction η of the potential difference lost in the needle lead is, according to (11, 3), determined by

$$\eta = \omega c_o r_o = \frac{c_o r_o}{\sqrt{LC}} \tag{12}$$

If—as above—$r_o = 200$ ohms, $c_o = 5$ cm. $= 0.55 \times 10^{-11}$ farad, and $\omega = 1 \times 10^6$ then $\eta = 0.0011$. This correction is also quite negligible in all cases where the frequency does not substantially exceed the above value.

Equations (11) and (12), on the other hand, show the importance of as small a value as possible of the resistance r_o in the needle lead; but if the suspension wire for this reason is made short and thick, the electrometer will be too insensitive.

This difficulty was overcome by the arrangement shown in Figure 5a. The suspension wire is a quartz fibre Q, and the electrical connection is affected thru the stout wire S carrying a Wollaston wire W. The core of the Wollaston wire is a 0.004 mm. thick platinum wire and the silver coating is removed only from the middle part of the wire and retained at both ends. Originally we made two hooks of the stout ends of the wire and used the hooks as connections to the wire S and to the needle N, respectively.

While this connecting arrangement gave a perfectly satisfactory results in the tests described in the above-mentioned publications,[9] later investigations have shown that it is not always

[9] See note 6 *a* and *b*.

reliable. High contact resistances may appear between the hook *H* and the wire *S* and between the hook *K* and the vertical aluminum wire *T* carrying the needle *N*, and finally between the wire *T* and the needle *N* itself. These resistances may vary erratically and introduce considerable errors in the measurements. Similar observations were made by Nancarrow and Cohen (see note 8), who eliminated this difficulty by using a thin metal wire for both suspension and connection wire for the needle. The said authors were satisfied with the results thereby obtained and find this electrometer method very convenient and superior to other methods in accuracy.

We were not, however, quite satisfied with the said suspension arrangement which had some drawbacks, partly owing to the fact that the elastic properties of a thin metal wire are not nearly as close to the ideal as those of a thin quartz wire. Using metal wire the zero position of the needle will vary more or less from throw to throw and the throws will not be absolutely constant. Using quartz wire, the zero position does not move and—as we shall see later on—the throws are exceedingly regular. We, therefore, rearranged the suspension arrangement as shown in Figures 5a and 7, where the quartz wire *Q* carries the needle while the Wollaston wire *W* makes the connection having its two stout ends *H′* and *K′* soldered to the connecting wire *S* and to the carrier wire *T* of the needle, respectively. The electrical and mechanical connection between *T* and the needle *N* itself—made of aluminum foil—is secured by using the design shown in Figure 7. With this arrangement perfectly satisfactory results have been obtained and we prefer this design to the one used by Nancarrow and Cohen, at least where accuracy is essential.

3. The Discharging Key

One more source of error has to be considered, namely, the losses in the discharging key *N* (see Figures 4 and 6). At the beginning, great difficulties were encountered here. We first tried an ordinary discharge-key, but the throws of the electrometer were too small—as small as 1-20th of the value to be expected—and very irregular. This, no doubt, was caused by the closing-spark at *b*. Numerous arrangements were tried in order to effect a good and sure contact instantaneously. We may here mention that we tried polished copper and steel hammers making contacts against polished copper or steel plates, the impact being made with great velocity. None of the various meth-

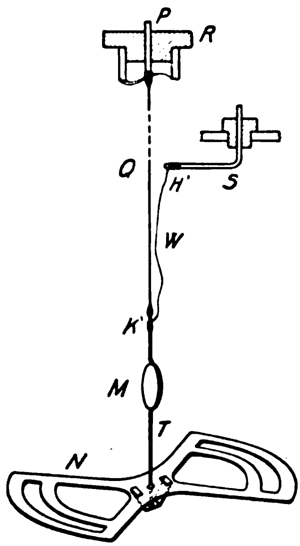


FIGURE 7—New Method for suspension of the needle and for providing electrical connection. (Compare Figure 5a.) In the tests described below, the quadrant electrometer had a quarts wire 90 mm. in length and 0.01 mm. thick. The Wollaston wire was about 50 mm. long and 0.004 mm. thick. The needle was made of aluminum foil 0.02 mm. thick.

ods where the contact was made between solid bodies gave the desired results. We next tried to make contact between a perfectly clean surface of mercury and a clean polished rod of metal with a rounded end (using steel, tungsten, or platinum) and large constant throws were thus obtained—provided that a perfectly clean mercury surface was prepared prior to each single discharge.

This leads to the conclusion that the contact should be made between volumes of mercury in vacuum, and this method when finally adopted gave perfectly satisfactory results. At the beginning we used the mercury lamp shown in Figure 8, where the contact is effected simply by tipping over the complete lamp, which causes a coherent stream of mercury to run from the one branch to the mercury in the other branch. In the following this key is called key I. In this key the contact-making jet is comparatively long and of course does not always have the same

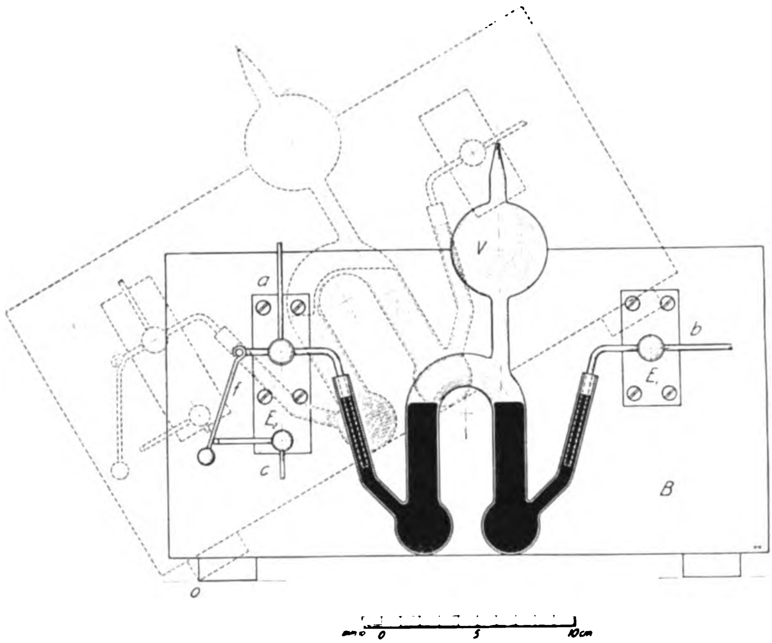


FIGURE 8—Key I. Mercury lamp used as discharging key. The connecting wires *a*, *b*, and *c*. corresponds to the connections denoted by the same letters in Figures 4 and 6. Using the arrangement of Figure 6, *a* and *c* are directly connected, thus shunting out the interrupter *f*. This is also the case with the keys shown in Figures 9-11. E_1 and E_2 are ebonite pieces. Similar lettering is used in Figures 9-11

cross-section. This may introduce an error of say 0.002 ohms while the total resistance of the key is about 0.016 ohms. In order to reduce this error we designed the key shown in Figure 9, having a very short jet.

The key is mounted on a wooden panel *B* revolving about the axis *O*. When the key is at rest, it takes the position shown in full lines, and in this position *b* is disconnected while *a* is in connection with *c*. Pushing the shaft *M* downward, the key is turned over to the position shown in broken lines, and the connection between *a* and *c* is broken before connection is made between *a* and *b* by bringing into contact the two volumes of mercury. When employing the arrangement Figure 6, *a* and *c* are directly connected and the interrupter *f*, thereby put out of operation.

A third type or key is shown in Figure 10.

If the handle *M* is brought up by a quick movement into the position shown in broken lines, the mercury drops from the bulb *V* down into the other end of the key, thus making contact between the two volumes of mercury—that is, contact is pro-

duced between a and b. The purpose of this key was to investigate whether it was of any importance to make the contact between the mercury volumes with great rapidity.

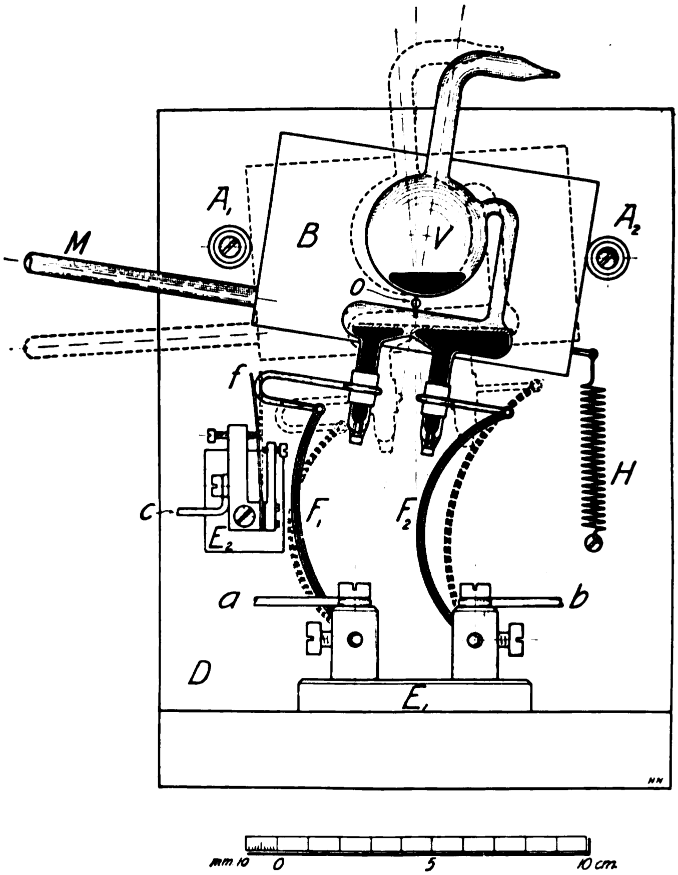


FIGURE 9—Key II. The mercury in the bulb V serves to regulate the volume of mercury in the lower, active part of the key

Beside these keys we have also tried one of the same design as was used by Nancarrow and Cohen[10] but mounted similarly to our own keys, as appears from Figure 11. This key was filled with hydrogen, while the keys I, II, and III, were evacuated to a pressure of not exceeding 0.0001 mm. of mercury.

With these four keys—and several others—a series of comparative measurements have been made, using the arrangement

[10] The key itself was delivered by A. C. Cossor, Ltd., London, and was the same type as used by Nancarrow and Cohen.

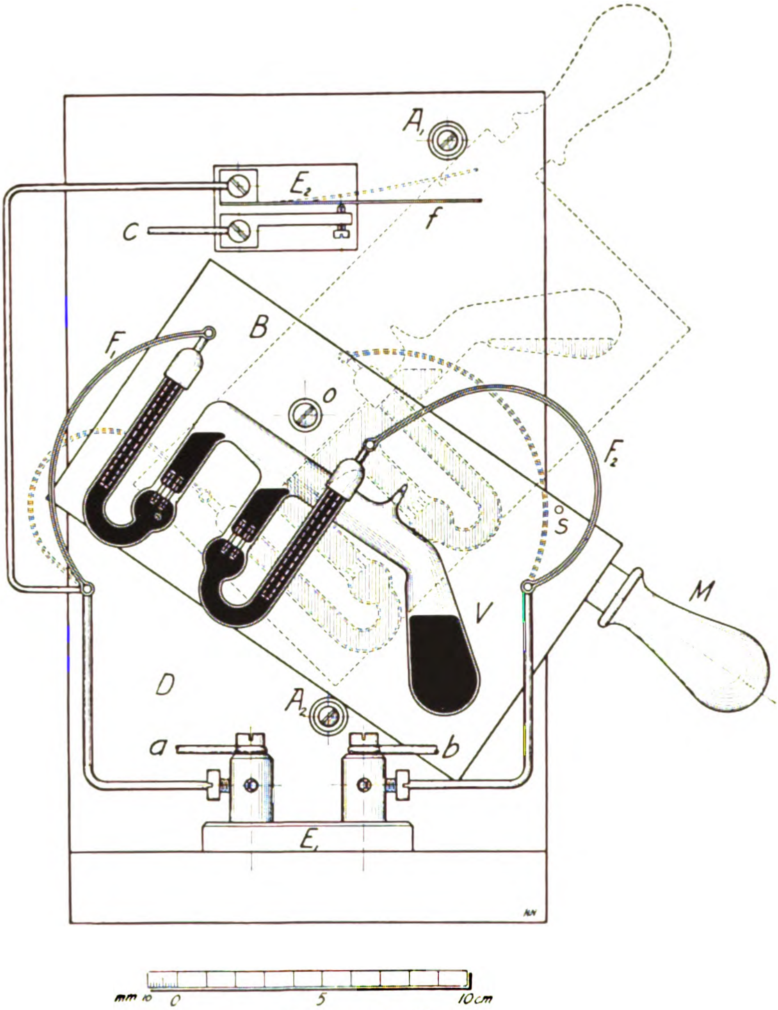


FIGURE 10—Key III

shown in Figure 4. With the same oscillating circuit and a constant value of V_o the throw of the quadrant electrometer for the various keys has been measured.[11] Ten readings have been taken for each of the keys, the mean value of which is shown in Table 1, where also is shown the mean error of the difference between the single throws and the mean value.

Column 4 of the table further shows the total resistance of the circuit ($R = R^o +$effective key resistance)—calculated on the

[11] The distance between mirror and scale was about 3 meters.

basis of the throw given in column 2. The resistance given for key II was determined in a special series of tests which will be mentioned below.

Column 5 shows the direct current resistance R' of the keys when closed. Column 6 shows the resistance $R°$+the "spark resistance" R_g of the key. We have accordingly $R°+R_g=R-R'$. Column 7 shows finally the difference between the "spark resistance" of key II and the same resistance of the other keys. Later on we shall see that this resistance for key II may, to a close approximation, be taken to be zero.

TABLE 1

COMPARISON OF VARIOUS KEYS. (V_o=450 VOLTS)

1	2	3	4	5	6	7
KEY	Throw Mean Value mm.	Mean Error %	Total Resistance of the circuit R Ohms	D. C. Resistance of the Key R' Ohms	The Circuit Resistance —"Spark Resistance," Calculated from Columns 4 and 5 $R_o+R_g=R-R'$ Ohms	"Spark Resistance"
I Fig. 8—Vacuum...	420.6	0.17	0.5099	0.016	0.4939	—0.0002
II Fig. 9—Vacuum...	416.7	0.047	0.5147	0.0210	0.4937	0
III Fig. 10—Vacuum...	422.4	0.058	0.5078	0.0132	0.4946	—0.0009
IV Fig. 11—Hydrogen (English key).	348.7	0.075	0.6151	0.0200	0.5951	—0.1014

Keys I, II, and III have practically no spark resistance while the hydrogen-filled, English key has a little over 0.1 ohm.

The table further shows that key II has the smallest mean error—less than ½ part per thousand. Least satisfactory in this respect is the key which was originally used—number I—but this, as mentioned before, is surely due to the relatively long "mercury thread" by means of which the connection is established between the two volumes of mercury.

Key II gives the most constant throws, has practically no spark resistance, and is exceedingly convenient to use. This type of key is no doubt the best one, and so far it has not shown any disadvantages whatever.

Besides these keys we have tried various others, and we have further tried various gas pressures in key II—using air as well as hydrogen. The result of these investigations may be summarized thus: As long as the mercury is perfectly clean, the spark resistance is very small, at least for pressures above 10 mm.

of mercury and below 0.1 mm. pressure. Between these two limits the spark resistance can, on the other hand, reach quite considerable values when, at the same time, the closing spark becomes very luminous.

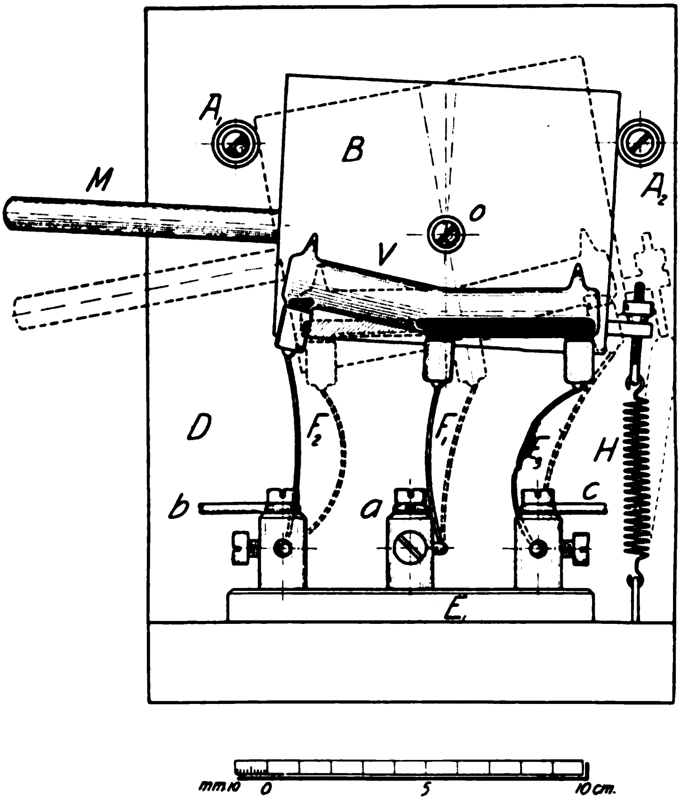


FIGURE 11—Key IV. Filled with Hydrogen and of same design as was used by Nancarrow and Cohen

If the key contains oxygen the mercury surface rapidly becomes dirty owing to oxidation produced by the contact sparks, and the result of this is a high and variable spark resistance. In this state, the key is useless. The key bulb must therefore be completely emptied of oxygen, either by pumping or by washing with hydrogen. However, keys containing hydrogen at considerable pressures show some spark resistance. The best way is to clean with hydrogen and then exhaust the bulb as far as possible. Treated in this manner the keys have—according to our experience—an unlimited life. In such a key, the closing spark for $V_0 = 400$

volts and $C = 25{,}000$ cm. is very faint and hardly visible in complete darkness.

Recently Professor J. T. McGregor Morris called my attention to a very interesting paper throwing much light on this question (J. T. Morris: "On Recording Transitory Electrical Phenomena by the Oscillograph," "The Electrician," June 7, 1907). In this paper it was shown that with the generally used contact-making devices, a permanent contact is not made instantaneously but only after several contacts have been made and broken. For an ordinary tapping key contact, this intermediate state lasts for about 0.023 seconds, for instance, with four intermediate makes and breaks. With a mercury pool and a needle point contact, the permanent contact is completed in from 0.002 to 0.001 of a second with one or two intermediate breaks.

This makes it perfectly clear why it is impossible to use the ordinary telegraph key as a discharge key in this method of measurement, and all the information given in the above-mentioned paper indicates that a contact-making device consisting of either a clean butt-ended metal rod and a clean mercury surface or between two clean mercury masses would make a permanent contact at once and could thus be used for the discharge key in question with good results.

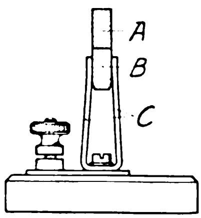


FIGURE A

It was also shown by the author of the above-mentioned paper that the intermediate breaks just after making the initial contact were due to rebounding of the contact pieces, and he overcame this difficulty—at least for the purpose he had in mind—by use of a modified knife blade contact of a form as shown in Figure A, corresponding to Figure 5 of cited paper. On the brass blade A of the switch an extension B of ebonite was fitted, having exactly the same width as the blade itself. To quote the paper: "As the ebonite extension was already in the jaws of the switch before the switch was tripped, when the brass entered the jaws of the switch it quickly glided in without any jar at the moment

of electrical contact. As far as the records taken with this form of switch contact go, it appears to work absolutely instantaneously, but certainly within 0.0001 of a second. Needless to remark, the contacts must be moderately clean, and an ordinary working pressure of jaws is necessary."

This indicates a possibility of using this modified knife blade contact—we call it type A key—for our discharge key and we therefore made such a key, and current oscillograms were taken for this key, for an ordinary telegraph key, and for the mercury-mercury key number II. In agreement with the above-mentioned paper, we found that the ordinary tapping key contact gave intermediate breaks while key A and the mercury-mercury contact gave smooth starting curves for the current, thus indicating that they made permanent contact at once.

We therefore tried key A as a discharge key in our method of measurement with the following results:

Key	Readings mm.	Remarks
Mercury-mercury	205, 205, 205, 205	
Key A	29, 81, 113, 59, 33 145, 48	Key closed directly by hand
	182, 188, 198, 168 201, 203, 202, 204, 203, 199, 204, 198, 202, 201, 200	Key closed by means of quick blow from a wooden hammer

It thus appears that the key A worked very rapidly was much better than the ordinary tapping key and, in some cases, came very close to the mercury key. But this latter key is far more constant and is much more convenient to use than key A, as it is necessary every now and then to grind the two sides of key A in order to get rid of small projections between the ebonite and the brass. This key also has to be kept very clean; otherwise it became far inferior to the mercury key. We therefore prefer the mercury key.

4. Carrying Out of Measurements

The method described above can be used in various ways for the determination of the radio frequency resistance of an oscillating circuit at its natural frequency, and consequently for determination of radio frequency resistance of inductances, condensers, etc., at any desired frequency.

We shall commence with a treatment of the first-mentioned problem. If the condenser of the oscillating circuit is practically

without leakage, then the arrangement shown in Figure 4 is employed. With the extra resistance r inserted in the oscillating circuit we have—as shown in part 1—

$$R+r=\frac{L}{2\beta P_r}V_o^2=\frac{K_o}{P_r} \tag{4,2}$$

where P_r is the throw of the quadrant electrometer with the total resistance $R+r$ in the circuit while K_0 is a constant as long as the potential V_o, the inductance L, and the ballistic constant β of the electrometer are kept constant.

Taking two measurements, one with $r=0$ and one with $r=r$, the resistance of the circuit itself is determined by

$$R=r\frac{P_r}{P_o-P_r}. \tag{5}$$

In order to check the accuracy of these measurements, various values of r may be inserted and the resistance R determined for each value by means of equation (5). In Table 2, there is recorded a series of such measurements.

TABLE 2

KEY II, DIAGRAM FIGURE 4, RESISTANCES IN OHMS

The added resistances $r=$	0.1	0.5	1.0	1.5	2.0	Mean value of R
R calculated by means of equation (5)	0.501	0.499	0.499	0.500	0.499	0.4996

Using the arrangement shown in Figure 6, where the condenser of the oscillating circuit is permanently connected to the battery during the measurements, the resistance of the circuit—as stated in paragraph 2—is increased by the resistance r^o determined by

$$r^o=\frac{1}{R_o}\cdot\frac{L}{C} \tag{8}$$

where R_o is the resistance of the conductor between the free terminal of the battery and the discharging key.

In Table 3, there is recorded an example of a series of such measurements taken on the same circuit as dealt with in Table 2.

TABLE 3

KEY II, DIAGRAM FIGURE 6, RESISTANCES IN OHMS

The added resistances $r=$	0.5	1.0	1.5	2.0	Mean value of $(R+r^o)$
$R+r^o$ calculated by means of equation (5) $=$	0.608	0.609	0.609	0.610	0.6090

According to Tables 2 and 3, the correction r^o should have the value of $0.6090-0.4996=0.1094$ ohms. The correction calculated by means of equation (8) is 0.1146 ohms ($L=1.34\times 10^{-3}$ hy., $C=1.11\times10^{-7}$ farad, $R=105{,}000$ ohms). The correction calculated is thus 0.0052 ohm greater than that measured. As stated in paragraph 2, the correction thus calculated is a little too large, and the agreement is therefore quite satisfactory.

The measurements recorded in Tables 2 and 3 were carried out with key II. For comparison there is recorded in table 4 a series of measurements on the same circuit but using key IV.

TABLE 4

KEY IV, DIAGRAM 4, RESISTANCES IN OHMS

The added resistance $r=$	0.1	0.3	0.5	1.0	1.5	2.0	Mean value of R
R calculated by means of equation (5) = ...	0.499	0.500	0.502	0.502	0.501	0.503	0.5011

Comparing with Table 2, we find that key II is a little better than key IV, but even the latter gives sufficient accuracy for practical use.

For these measurements there was used as the "added resistance" the rheostat shown in Figure 12, which is variable in steps of 0.1 ohm from 0.1 to 2.0 ohms. The resistance is made of a single constantan wire provided with taps soldered on at suitable intervals.

The exact adjustment of the individual values was done by carefully scraping away some of the wire.

For measuring radio frequency resistances above 5 ohms, we used as the added resistance radio frequency rheostats of higher resistances. This method may conveniently be used for measuring the radio frequency resistances of coils up to several hundred ohms.

In carrying out the measurements on highly damped coils, that is, coils for which R/L has a high value, it will often be preferable to insert another coil (L_1) into the oscillating circuit, (L_1) having a great inductance and a small but known radio frequency resistance. The throws of the electrometer may thereby be considerably increased.

To measure an unknown radio frequency resistance x having no inductance or capacity, we simply insert this resistance instead of the known resistance r in Figure 4. If beforehand we

have determined in the way set forth above, the resistance R of the circuit itself, then, by means of equation (4, 2), we may find the constant $K_o = P_o R$. We then have

$$x = \frac{K_o}{P_x} - R \qquad (13)$$

where P_x is the throw of the electrometer after having inserted the resistance x in the circuit.

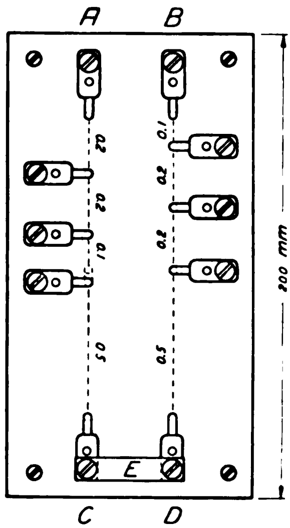


FIGURE 12—Rheostat used as "added resistance" in these measurements and made of a single 0.3 mm. Constantan Wire

To measure the difference in radio frequency resistance of various condensers having equal capacity, we successively insert various condensers into the oscillating circuit and determine the corresponding throws $P_{c(n)}$ of the electrometer. Having previously determined the constant K_o we then have

$$R_{c(n)} = \frac{K_o}{P_{c(n)}} \qquad (14)$$

The condenser, denoted by (n), has then an effective resistance at the applied frequency of $(R_{c(n)} - R_{c(m)})$ ohms more than the resistance of the condenser (m).

If one of the condensers is to be considered as leakage-free, and if we denote this by (o), then the resistance $r\ (n)$ of the condenser (n) is given by

$$r_n = R_{c(n)} - R_{c(o)} \qquad (15)$$

We have found that carefully made condensers, using pure ruby-colored mica as dielectric, and with plates made of not too thin copper foil, with a perfectly clean surface, do not show any more loss in dry air than an air condenser of the same capacity.[12] Such mica condensers may, therefore, be considered as being without loss. But it is very essential that the air be dry, for if this is not the case, the mica condensers will show a considerable loss. It is therefore necessary to embed the condensers in paraffine or to use some other suitable means of keeping them dry.

In the following Table 5, giving the results of measurements of radio frequency resistance of some condensers, condenser number (o) is considered free of loss.

TABLE 5

COMPARISON OF THE EFFECTIVE RESISTANCES OF DIFFERENT CONDENSERS AT FREQUENCY OF ABOUT 15,000. ($C = 100{,}000$ cm.)

Index (n)	Throw $P_{c(n)}$ mm.	Total Resistance of Circuit $R_{c(n)}$ Ohms	Resistance of Condenser R_n ohms	Dielectric Material		Thickness of Copper Foil mm.	Remarks
				Material	Thickness mm.		
0	437.7	0.4735	0.0000	Ruby-colored Mica	0.110	0.06	Copper foil carefully cleaned of grease and oil
1	399.4	0.5189	0.0454	"	0.112	0.06	Copper foil not cleaned
2	412.5	0.5000	0.0265	"	0.200	0.05	Preparation unknown
3	406.7	0.5071	0.0336	Green-colored Mica	0.135	0.06	Copper foil carefully cleaned of grease and oil
4	96.5	2.1373	1.6638	Celluloid	0.117	0.06	

If loss-free condensers or condensers of a known loss are used the method here described is very convenient for the determination of radio frequency resistances of any coil or condenser at any desired frequency, but it is hardly necessary to dwell any longer upon this point. We will therefore consider here only one application of the method, namely the determination of radio frequency resistances of dielectric materials. The arrangement shown in Figure 13 is used for this purpose. In parallel to the variable condenser C is placed an air condenser C'. Between the plates of the latter is placed a sheet (p) of the material to be investigated. The thickness of this sheet is denoted by d cm. If the

[12] Air condensers will generally show even a little higher loss owing to longer connections. It is very essential that the different condensers should be treated alike, especially with regard to temperature and humidity of the air. By breathing on a condenser one may materially increase its effective resistance

The described method of measurement is so sensitive that the influence of variations in humidity and temperature of the air on the effective resistance of oscillating circuits is easily observed. The throws of the galvanometer will, for example, generally have different value when the door of the test room is open from the value when it is closed.

material shows any perceptible conductivity for direct current, it is appropriate to place two very thin mica sheets e (0.02 – 0.05 mm.) between the sheet p and the condenser plates as shown in Figure 13. This also makes it unnecessary to have the battery permanently connected to the condenser and renders needless the correction mentioned in paragraph 2.[13]

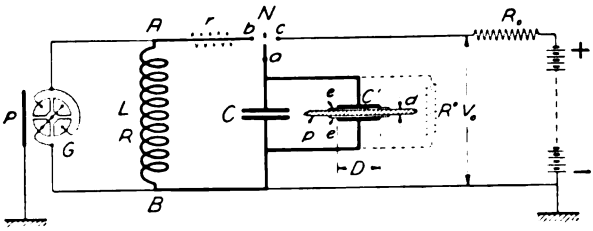


FIGURE 13—Arrangement Used for Determination of Radio Frequency Resistances of Dielectric Materials

The measurements are carried out as follows: The radio frequency resistance of the oscillating circuit without the test sheet in C' is determined in the usual manner by $R = r\dfrac{P_r}{P_o - P_r}$.
Next the sheet p is introduced into C' and the capacity of C is reduced until the total capacity has the same value as before. If the throw of the electrometer is now P', then the increase of the radio frequency resistance—R'—caused by the sheet p is determined by

$$R' = R\frac{P_o - P'}{P'} = r\frac{P_r}{P_o - P_r} \cdot \frac{P_o - P'}{P'} \tag{16}$$

This increase of resistance may be imagined, for instance, as a resistance—R^o—shunted across the condensers C and C' and, according to formula (6) or (8), paragraph 2, R^o is determined by

$$R^o = \frac{1}{R'} \cdot \frac{L}{C_1}, \tag{17}$$

where C_1 is the total capacity.

[13] The introduction of these mica-sheets is, however, permissible only if the following relation is satisfied:

$$\frac{n\varepsilon\rho d}{t} \cdot \frac{1}{18\times10^{11}} \gg 1,$$

where n is the frequency, ε the dielectric constant of the mica, t the total thickness of the mica sheets (cm.) and ρ the specific resistance of the dielectric material while d is the thickness of the sheet p. If the above relation is not satisfied, a considerable part of the potential drop across the condenser C' will be found across the mica sheets. This introduces an error which, under unfavorable conditions, may be very considerable. In such cases the arrangement shown in Figure 6 must be applied.

If the thickness d of the sheet is small compared with the diameter D of the plates in the air condenser C', then the resistivity ρ of the test plate p at the applied frequency is determined with sufficient accuracy by

$$\rho = R^o \frac{F}{d} \text{ [Ohm per cm}^3\text{]} \tag{18}$$

where $F = \frac{\pi}{4} D^2$ is the area of the air condenser.

5. Determination of the Ballistic Constant of the Electrometer

In the investigations treated above, there is an uncertainty in one regard. In table 1, paragraph 3, we have arbitrarily put the "spark resistance" of key II as equal to zero. The measurements so far considered, give us only the difference between the spark resistances of the various keys. The fact that, in spite of considerable differences in design, the three evacuated keys show very little difference in "spark resistance" makes it probable that this resistance is very small in all of them—but it is not a proof. Such proof can be produced only by an independent determination of the ballistic constant of the electrometer.

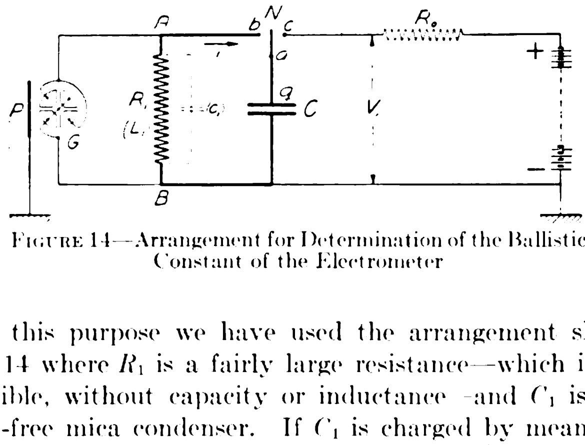


Figure 14—Arrangement for Determination of the Ballistic Constant of the Electrometer

For this purpose we have used the arrangement shown in Figure 14 where R_1 is a fairly large resistance—which is, as far as possible, without capacity or inductance—and C_1 is a good leakage-free mica condenser. If C_1 is charged by means of the key N to the potential difference, V_1, and then discharged thru the resistance R_1 by throwing the key N over into the position $a\ b$, the potential difference between the points A and B will vary during the discharge. The time integral (B) of this difference squared is easily calculated beforehand.

In order to evaluate the influence of a small inductance in R_1, we will carry out the calculations under the assumption that,

beside the resistance R_1, there is an inductance L_1, assuming the circuit C, L_1, R_1 to be aperiodic.

If we denote the initial value of the potential difference across the condenser by V_o, and introduce the following abbreviations

$$\kappa_1 = \frac{R_1}{2 L_1} \text{ and } h_1{}^2 = \frac{1}{L_1 C}.$$

then we have

$$v_1 = \frac{V_1}{2\sqrt{\kappa_1{}^2 - h_1{}^2}} \left\{ \begin{array}{l} (\kappa_1 + \sqrt{\kappa_1{}^2 - h_1{}^2})\, \varepsilon^{(-\kappa_1 + \sqrt{\kappa_1{}^2 - h_1{}^2})t} \\ + (-\kappa_1 + \sqrt{x_1{}^2 + h_1{}^2})\, \varepsilon^{-(\kappa_1 + \sqrt{\kappa_1{}^2 + h_1{}^2})t} \end{array} \right\} \tag{19}$$

from this we get

$$B = \int_o^\infty v_1{}^2\, dt = \frac{1}{4} V_1{}^2 \frac{4\kappa_1{}^2 + h_1{}^2}{\kappa_1 h_1{}^2} = \frac{1}{2} V_1{}^2 \left(R_1 C + \frac{L_1}{R_1} \right). \tag{20}$$

Using the arrangement of Figure 14, R_1 will generally not be less than 10,000 ohms and C generally not less than 0.1×10^{-6} farad. The last term inside the parenthesis in equation (20) is then negligible in comparison with the first, and we may therefore, to a close approximation, put

$$B = \frac{1}{2} V^2{}_1 R_1 C \tag{20, 1}$$

If the resistance R_1 has a capacity C_1, as indicated by the broken lines of Figure 14 (but without resistance in the connecting wires), this capacity will instantaneously be charged at the beginning of the discharge to the potential difference $V_1 \cdot \frac{C}{C + c_1}$ C also will have this potential. In this case, we get

$$B_1 = \frac{1}{2} \left(V_1 \frac{C}{C + c_1} \right)^2 \cdot R_1 (C + c_1) = \frac{1}{2} V_1{}^2 \frac{C^2}{C + c_1} R_1 \tag{22}$$

From this it follows, by comparison with (20, 1),

$$\frac{B_1}{B} = \frac{C}{c_1 + C} \tag{23}$$

For good resistances, this correction also is quite negligible.

Finally we have to consider the influence of the loss which is caused in this case by the making of the contact at b. Here we find, just as is set forth above, that if an ordinary discharge key is applied at N, the throws of the electrometer G are too small and somewhat irregular. Even if the conditions are here far less unfavorable than when discharging thru the oscillating circuits of Figures 1 and 4, yet such a key is inapplicable. Using key II,

the throws are perfectly constant, and as only a single closing of the current is effected, we can surely consider in this case the resistance of the key to be negligible in comparison with the large resistance R_1. We can therefore without hesitation apply the arrangement of Figure 14 to the determination of the ballistic constant. All that is required is to calculate the value B by means of equation (20, 1) and to measure the throw P of the electrometer; we then have

$$B = \beta P, \tag{24}$$

where β is the ballistic constant.

Having in this manner determined the value of β we can, by means of formula (4, 1) in paragraph 1, namely

$$R = \frac{L}{2\beta P_o} \cdot V_o^2 \tag{24, 1}$$

determine the resistance R of the circuit.

In this manner the resistance of a certain circuit was determined to be 0.684 ohms while a determination according to the method using added resistance described in paragraph 4 gave a resistance of 0.683 ohms—an agreement which is quite satisfactory. We can therefore, without hesitation, put the resistance of key II as equal to zero.

6. Concluding Remarks

The method offers among others the following advantages:

(1) Much greater accuracy than the methods hitherto applied, especially for feebly damped circuits.

(2) The method is simple, convenient, and quick.

(3) No special generator for radio frequency oscillations is required, and there is therefore no complication arising from tuning or coupling.

As a slight disadvantage, it may be mentioned that the quadrant electrometer requires a fairly stable support, while on the other hand it is safely transportable with the clamping device shown in Figures 5*a* and *b*.

In working out this method I have had excellent assistance from Mr. J. P. Christensen, Mr. Chr. Nyholm, Mr. B. B. Rud, Mr. Kay. Christiansen, Mr. Nørregaard, and Mr. J. Egelund-Nielsen.

Part of the expenses have been defrayed from a grant received from the "H. C. Ørsteds Fond for teknisk-videnskabelig Forskning," founded by The Great Northern Telegraph Company.

Telegraph and Telephone Laboratory
of The Royal Technical College,
Copenhagen, September, 1924.

SUMMARY: The author criticizes the existing methods for the measurement of radio frequency resistance and describes a new electrometer method, where a quadrant electrometer is put across the inductance in an oscillatory circuit. The condenser of this circuit is charged to the voltage V_o and discharged thru the inductance by means of a special key.

The throw of the electrometer will then be proportional to $V_o^2 - L/2R$. The theory of the method is given and it is shown how to eliminate the different possible sources of error. With this method the radio frequency resistance of even a very feebly damped circuit may be determined with an error well within one part in a thousand, and this determination may be made in a few seconds.

NOTE ON TELEPHONE RECEIVER IMPEDANCE*

By

E. Z. Stowell

(University of Nebraska)

Some time ago the writer made a series of measurements of telephone receiver impedance to radio frequencies at the Bureau of Standards Radio Laboratory. It is believed that the results will be of interest.

The method employed was that of resistance variation, using the circuit of Figure 1. In the circuit L is a standard inductor, a single-layer solenoid on a polygonal frame ; C is a variable standard air condenser, with a vernier condenser in parallel; R is a pure resistance which can be discontinuously varied; a thermocouple T and a galvanometer G for direct current indicated resonance. The entire circuit was enclosed in a grounded copper cage, with the exception of the galvanometer. The leads to the

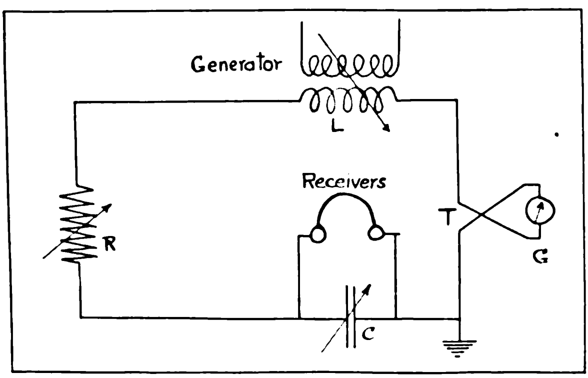


Figure 1

galvanometer were enclosed in grounded lead sheath. On the side of the cage next the coil the screen was widened sufficiently to permit of the introduction of an emf. from the output coil of a 250-watt electron tube oscillator.

The resistance of the circuit was determined as follows:

*Received by the Editor, October 24, 1924.

The source and the measuring circuit were both tuned to the desired frequency and the galvanometer reading noted. Then one of the standard resistors R was introduced into the circuit from outside the cage. The circuit remained in tune, and the galvanometer was again read. Assuming (1) that the applied emf. was constant, and (2) that the galvanometer deflections were proportional to the square of the current, the resistance of the circuit originally was

$$r = \frac{\Delta r}{\sqrt{\frac{d_1}{d_2}} - 1}$$

where Δr is the extra resistance introduced by the standard resistor, d_1 and d_2 being the deflections before and after the introduction of Δr.

Suppose the resistance of the measuring circuit to be measured in this manner. The telephone receivers, suspended horizontally in an auxiliary grounded cage, are introduced in parallel with condenser C from without. In general the telephones not only add resistance to the circuit, but detune it as well. The circuit is again tuned by the condenser, the change in capacity being noted and the resistance again measured as described above. The resistance and reactance of the inserted phones are now known from these relations:

$$\left.\begin{aligned} R &= \frac{\Delta}{\left(\frac{\delta}{C_1}\right)^2 + (\Delta C_2 \omega)^2} \\ X &= \frac{1}{C_2 \omega}\left[1 + \frac{\frac{\delta}{C_1}}{\left(\frac{\delta}{C_1}\right)^2 + (\Delta C_2 \omega)^2}\right] \end{aligned}\right\} \begin{aligned} \Delta &= R_2 - R_1 \\ \delta &= C_2 - C_1 \end{aligned}$$

where C_1 and R_1 are the capacity and resistance in the circuit without the headset, the insertion of which changes them to C_2 and R_2, respectively. C_1 and C_2 of course include the capacity of the inductor L. ω is 2π times the applied frequency.

Observations were made on about 50 pairs of telephones, comprising 14 makes, at 8 isolated frequencies between 6,000 cycles sec. and 1,000,000 cycles sec., Mr. C. T. Zahn collaborating. The results for a typical pair of telephones now on the market are given in Figure 2.

It is seen that a pair of receivers behaves electrically like a wave trap, or parallel resonance circuit. The receivers have an electrical natural frequency quite distinct from the natural

frequency of the diaphragm ; in the case shown, it occurs at 12,000 cycles/sec. At this frequency, the resistance suffers an enormous increase. Above this frequency, the telephones behave like a condenser with rather large dielectric loss.

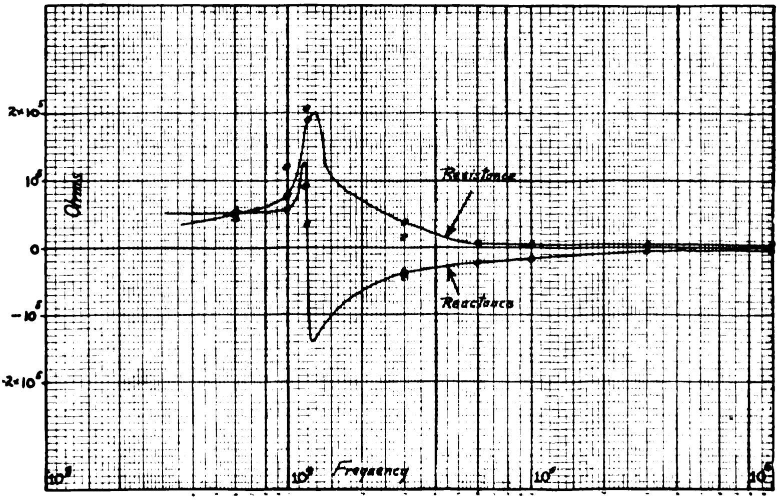


FIGURE 2

It appears that the receivers have some characteristics in common.

1. Their natural frequency falls in the range from 9,000 to 15,000 cycles/sec. with occasional exceptions.
2. Their resistance at the natural frequency is of the order 100,000 to 200,000 ohms.
3. At frequencies above 50,000 cycles/sec. the telephones may be considered a condenser of the order of 10^{-10} farad.
4. Their resistance above 50,000 cycles/sec. is inversely proportional to frequency and may be calculated from the relation

$$R = \frac{\psi}{C\omega}$$

where ψ is the constant phase difference and C the capacity of the phones. ψ varies with the make from 0.04 to 0.10 ; 0.06 is an average value.

5. At frequencies of the order of 1,000,000 cycles/sec., receivers differ very little among themselves, and little change is produced in the impedance by removing the electromagnets, indicating that the capacity of the

leads furnishes the major portion of their impedance. At these frequencies all the receivers examined act like condensers of capacity very nearly 10^{-10} farad.

With regard to the precision involved in these measurements: the ratio of probable error of the impedance to the impedance itself is about 0.04 or 4 percent at low frequencies; at resonance about 10 percent for the resistance and 50 percent for the reactance, due to the rapid change in the latter; above resonance, about 5 per cent and decreasing as the frequency mounts.

The data for any given pair of telephones, as for example those of Figure 2, can be explained fairly satisfactorily by assuming a symbolic circuit for them as in Figure 3. Here L is

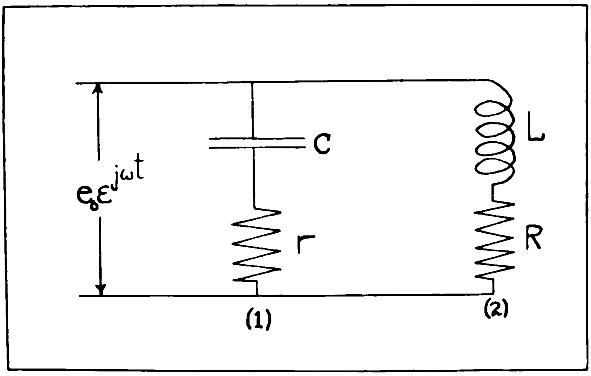


FIGURE 3

the direct current inductance of the electromagnets of the order of a few henrys ; C is the capacity of the leads of the order of 10^{-10} farad; R the resistance of the electromagnets which increases in a complicated manner with the frequency, and r is the resistance of condenser C, and inversely proportional to frequency. It can be shown that if these four quantities are known, the resistance and reactance of the telephones at any angular velocity ω are

$$R = \frac{r\,(R^2+L^2\,\omega^2)+R\left(r^2+\frac{1}{C^2\,\omega^2}\right)}{(R+r)^2+\left(L\,\omega-\frac{1}{C\,\omega}\right)^2}$$

and

$$X = \frac{L\,\omega\left(r^2+\frac{1}{C^2\,\omega^2}\right)-\frac{1}{C\,\omega}\,(R^2+L^2\,\omega^2)}{(R+r)^2+\left(L\,\omega-\frac{1}{C\,\omega}\right)^2}$$

In Figure 2 the crosses are values computed from these formulas; where none appear, they fall so close to the observed points that they were omitted for clearness.

Acknowledgment is due Director Burgess of the Bureau of Standards for his courtesy in allowing publication of this material.

Brace Laboratory of Physics,
University of Nebraska.

DISCUSSION* ON "ON THE RADIATION RESISTANCE OF A SINGLE VERTICAL ANTENNA AT WAVE LENGTHS BELOW THE FUNDAMENTAL"

AND

"ON THE OPTIMUM TRANSMITTING WAVE LENGTH FOR A VERTICAL ANTENNA OVER PERFECT EARTH"

BY STUART BALLATINE

BY

BALTH. VAN DER POL, D.SC. (MEMBER)

(PHYSICIST, PHILIPS' INCANDESCENT LAMP WORKS, EINDHOVEN, HOLLAND)

In the two above-mentioned papers, which appeared on pages 823 and 833, respectively, of Volume 12 of these PROCEEDINGS, the writer first calculates the radiation resistance of a vertical antenna loaded at the bottom only so that a current distribution over the antenna results with a current node at the top. The writer finds (formula (19) page 829), the radiation resistance to be a function of a single variable $a = \frac{2\pi l}{\lambda}$, this variable being determined by the ratio of the antenna length l to the wave length λ. The function is composed of the well-known integral sine and integral cosine functions. Further it is shown by the writer, that, when the antenna is excited at a wave length below the fundamental, the vertical distribution of the radiated energy, as represented by polar diagrams, is in certain circumstances markedly different from the distribution obtained from an antenna oscillating at its fundamental wave length. Finally it is suggested that important information concerning upper atmospheric reflection may be obtained by exciting an antenna in such a way that no energy is radiated in a horizontal direction, but at an elevated angle only. Any energy thus reaching a distant receiver must then be necessarily reflected and thus important data concerning the "Heaviside layer" might be obtained.

I may be allowed to point out that seven years ago I investi-

*Received by the Editor, January 2, 1925.

gated exactly the same question. The results of this research were published in the "Proceedings of the Physical Society of London," XXIX, page 269 (1917), and in the "Jahrbuch für drahtl. Telegraphie," XIII, page 217 (1917). In my paper, however, the problem was tackled in a somewhat more general way than in Dr. Ballantine's investigations, loading of the antenna at the top as well as at the bottom being considered. The radiation resistance, instead of being a function of the ratio of the antenna length to the wave length only, was found also to depend upon the boundary conditions at the top of the antenna, it being different, depending on whether a capacity at the top, in the form of a horizontal wire, is present or not.

In fact Dr. Ballantine's formula (19), page 829, for the radiation resistance in ohms:

$$R_{loop} = 60\left[\cos^2 a \cdot S_1(2a) - \frac{1}{4}\cos 2a \quad S_1(4a) - \frac{1}{2}\sin 2a\left\{S_1(2a) - \frac{1}{2}S_1(4a)\right\}\right] \tag{1}$$

is a special case of my more general formula (29), on page 279 of the "Phys. Soc. Proceedings (London)," which in the present notation reads

$$R_{loop} = 60\left[\sin^2\eta \cdot S_1(2a) + \frac{1}{4}\cos 2\eta \cdot S_1(4a) + \frac{1}{2}\sin 2\eta\left\{S_2(2a) - \frac{1}{2}S_2(4a)\right\} + \frac{\cos^2\zeta}{2}\left(\frac{\sin 2a}{2a} - 1\right)\right] \tag{2}$$

where:

$$\cos\zeta = \frac{I_{top}}{I_{loop}}$$

$$\cos\eta = \frac{I_{bottom}}{I_{loop}}$$

$$\eta + \zeta = a = \frac{2\pi l}{\lambda}$$

$$S_1(x) = \log x + 0.577216 - Ci(x)$$

$$S_2(x) = Si(x).$$

For, with $\frac{I_{top}}{I_{loop}} = 0$, $\zeta = \frac{\pi}{2}$ and $\eta = a - \frac{\pi}{2}$, and (2) is reduced to (1).

In my paper the vertical distribution of the radiated energy is also fully considered and depicted in seven drawings. It was also suggested there that, in order to get information about the higher atmosphere, it was feasible to send waves upwards at an inclined angle by exciting an antenna at a higher harmonic.

Recently very important experiments along these lines have been carried out,[1] in England, France, and Germany, and, up to the present, cases have been reported where at great distances better reception was obtained when the transmitting aerial was excited at a higher harmonic than when it was oscillating at the fundamental frequency.

Stuart Ballantine (by letter): I wish to thank Dr. van der Pol for calling our attention to his important paper, and for his courtesy in sending me privately copies of it and of the foregoing discussion. Upon examination I find that he has covered practically the ground of my first paper and also independently suggested the usefulness of the specially-excited antenna for Heaviside layer experiments. His mathematical exposition, addressed to mathematical-physicists, is somewhat more prolix than my own (in the ratio of forty-eight to seventeen lines), and is unfortunately not carried to a numerical conclusion, which, of course, is the part of real interest to most of us. I remember having spent about three days deriving the formulas (19) for the radiation resistance, whereas the numerical computations for Figure 3 consumed nearly three weeks. The sine-integral and cosine-integral tables were not complete and I had to build them up with the aid of the asymptotic formula. As to the second paper, on the optimum transmitting wave length of the ideal vertical antenna, I do not find any references in Dr. van der Pol's article to this subject. In view of this I do not now think that the publication of my papers is such an unfortunate duplication as I had been obliged to think after first glancing thru Dr. van der Pol's paper; nevertheless, I must express regret at not having sooner discovered his work, in order that I could have added a reference to it commensurate with its importance and priority.

The superior generality which Dr. van der Pol claims for his investigation seems to me of more academic than practical significance. I see no way of making the current at the free-end of the antenna different from zero which does not involve the connection of additional structure. If this is done, of course, the problem is changed and becomes that of a complex antenna. There seems to be a mistake in this part of Dr. van der Pol's paper, for he adds a horizontal section without taking any furher account of it than its effect on the current distribution over

[1] R. Mesny, "l'Onde électrique," III, page 99 (1924).
A. Meisner, "Jahrb. d. drahtl. Tel. u. Tel.," XXIV, page 884 (1924).

the vertical portion. Over a perfectly conducting plane earth the horizontal part will contribute only a horizontal component of electric force, so that as far as a vertical receiving antenna on the horizon is concerned it is sufficient to consider only the vertical field produced by the vertical section with its altered current distribution. But if the radiated power is to be calculated, it is necessary to know the field intensities in all points of the surrounding surface and the contribution of the top section to these intensities must be computed. Also the radiations of the two sections cannot be considered separately since the principle of superposition fails for the energy. Dr. van der Pol does not specify that the finite current at the end of the vertical antenna is due to a top section, but he vaguely refers to such a section and gives in formula (31) an expression for the radiation resistance "*of a flat-top antenna without any load at the bottom.*" If he really means this, the formula is, of course, inadequate, and if he does not, it is an abstract generalization. For the same reason only three of the energy distribution diagrams given seem of actual importance. These are Figures 1, 8, and 10. Figure 8 corresponds to Figure 3 of my second paper, characterizing the special Heaviside-layer transmitter.

It should be mentioned that the possibility of eliminating the direct radiation along the earth is not confined to a special excitation of the vertical antenna; a horizontal section can be added provided the earth be of sufficiently good conductivity. In this case a current distribution in the vertical part of the type shown in Dr. van der Pol's Figure 6 (reproduced below), would produce no vertical electric force on the horizon. Over earth of imperfect conductivity this would not be true, for as H. von Hoerschlemann[1] has shown, the horizontal section is then capable of producing a vertical electric force on the horizon by the induction in the earth of vertical current components which are not symmetrical about the lead-in.

I had read in "l'Onde électrique," after my paper had been submitted for publication, some fragmentary accounts of the preliminary experiments of Commandant Mesny. I hope that this work will be systematically and diligently pursued. In this connection I may say that my colleague, Mr. R. W. Seabury, of Boonton, New Jersey, is erecting at the seashore a special transmitter of the type described in the last section of my second paper, with which we hope to be able to make some careful experiments over sea-water.

[1] H. von Hoerschlemann: "Jahr. d. draht. Teleg.," volume 5, page 14 (1912).

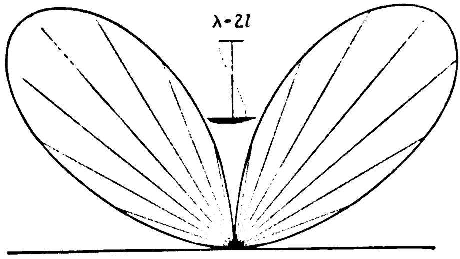


FIGURE 6—Reproduced from Dr. van der Pol's paper in the "Proceedings of the Physical Society of London," 24, 269

It was discovered some time ago by American amateurs experimenting with unusually short wave lengths that the transmission of certain short waves during the day was surprisingly good. This has been popularly attributed to the wave lengths employed, due I suppose to some esoteric predilection of the Heaviside reflecting mechanism for these wave lengths. I believe that a better explanation can be based on the probability that most of these short waves were produced by antenna systems which had previously been designed to operate at wave lengths four times as long, that is to say, fifty-meter waves were being produced by antennas, the fundamentals of which were in the neighborhood of two-hundred meters. In this case the energy would be radiated up into the air and if, according to modern speculation, the Heaviside-layer is a better reflector by day than by night, the superior long-distance transmission could be readily explained. It would thus seem that the optimum transmitting wave length, while being definitely indicated in normal circumstances, would in the presence of celestial reflection depend upon many things, the height of the reflector, the distance of the receiver, and so on, and would therefore dodge specification. For perfect earth it is 0.39 of the fundamental (idealized case considered in my paper); for ordinary earth it approaches the half-fundamental (as I shall show in a forthcoming paper); and for Heaviside-layer transmission it could be anything but would most likely be somewhere near the quarter-fundamental. All of these wave lengths are substantially below the fundamental, in a region of which I believe we have been too neglectful.

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

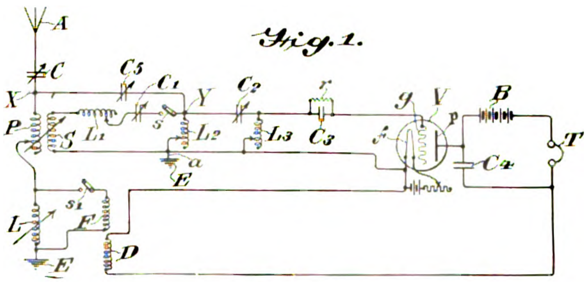
Issued January 6, 1925–February 24, 1925

By

John B. Brady

(Patent Lawyer, Ouray Building, Washington, D. C.)

1,521,777—D. G. McCaa, filed July 10, 1923, issued January 6, 1925. Assigned to the Electric Apparatus Company, Parkersburg, Pennsylvania, a corporation of Pennsylvania.



Number 1,521,777—Radio System

Radio System, having a circuit arrangement for eliminating interference or other disturbances. The frequency of the energy which it is desired should be excluded or reduced in effect and represented by oscillations set up by static, atmospherics or other natural electricity, is caused to differ from the frequency of the energy which it is desired shall be received, and the undesired oscillations are neutralized by an opposing and substantially equal potential caused by the received undesired oscillations, leaving the desired oscillations to effect the desired signals with substantially no disturbance by the undesired oscillations. When the undesired radio frequency oscillations are artificially produced as by transmitters emitting oscillations of frequency differing from the desired oscillations, their disturbing effects are similarly eliminated or reduced. The desired oscillations, prefer-

*Received by the Editor, March 7, 1925.

ably first amplified, are utilized to change the reactance of a circuit or path in which both the desired and undesired signals are received to materially increase the ratio of amplification of the undesired oscillations to the amplitude of the undesired oscillations. Similarly, with regard to undesired oscillations artificially produced and having a frequency different from the desired oscillations, the desired oscillations, preferably first amplified, effect a change of reactance of a circuit in which both the desired and undesired oscillations are received to render the circuit or path resonant to the desired oscillations and de-tuned or non-resonant with respect to the undesired oscillations. A local source of radio frequency oscillations, having a frequency differing from the desired oscillations, is utilized to effect a variation of reactance whereby the desired oscillations are caused to partake of maximum amplitude in a certain part or branch of said circuit of path and thereby increase the ratio of the effect of the desired oscillations to the effect of the undesired oscillations in the ultimate signal-translating instrument.

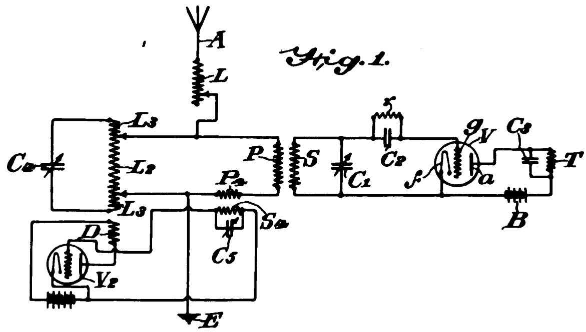
1,522,020—A. Maurer, filed January 29, 1924, issued January 6, 1925.

TABLE FOR RADIO APPARATUS, in which a loop is pivotally mounted beneath the table top and arranged to be operated by an extended shaft from the front of the table. The shaft has a rotatable dial on the extremity thereof and is gear-connected with the loop so that rotation of the shaft operates to position the loop in the desired plane.

1,522,070—T. H. Nakken, filed November 3, 1920, issued January 6, 1925. Assigned to Naamlooze Vennootschap Nederlandsche Lumingtron Maatschappli, Rotterdam, a Company of Netherlands.

MEANS FOR TRANSFORMING LIGHT IMPULSES INTO ELECTRIC CURRENT IMPULSES, by means of electron emission from a photoelectric body. The photoelectric body is enclosed within a tube exhausted of air. A cathode is provided in the tube for emanating electrons. The photoelectric body is subjected to the action of a variable exterior source of light and is provided in an electric circuit in which the intensity of current will be varied in accordance with the variations in the source of light. The tube is provided with a plate and a control electrode or grid whereby the exposure of the photoelectric body to the electron discharge may be controlled.

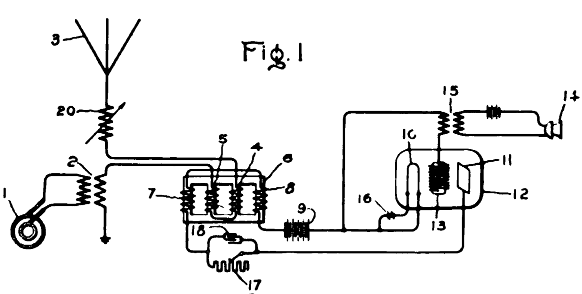
1,522,136—D. G. McCaa, filed July 29, 1924, issued January 6, 1925. Assigned to The Electric Apparatus Company, Parkersburg, a corporation of Pennsylvania.



Number 1,522,136—Receiving System

Receiving System, having a means for eliminating or reducing the effects of electrical disturbances in the reception of signals. The desired signal and the disturbing effects are divided between reactive paths, one of which is employed for effecting the translation of the desired signals, and with another of which is associated means for impressing thereon a part of the energy of the desired signals previously amplified, to cause a change of reactance, and thereby withholding from the signal-translating path the effects of the undesired oscillations to greater degree than the effects of the signal-representing or desired oscillations.

1,522,221—E. F. W. Alexanderson, filed November 20, 1914, issued January 6, 1925. Assigned to General Electric Company, a Corporation of New York.



Number 1,522,221—Method of and Means for Controlling Alternating Currents

METHOD OF AND MEANS FOR CONTROLLING ALTERNATING CURRENTS in transmitting systems for radio telephony, wherein large amounts of energy may be controlled by the small current variations produced by sound waves in an ordinary telephone transmitter, in such a way that the sound waves may be faithfully reproduced in suitable receiving apparatus at a distant point. The patent describes the Alexanderson magnetic amplifier. A reactance is connected in the antenna circuit. A second circuit including a magnetizing winding is inductively related to the reactance in such a manner that the impedance of the antenna circuit may be varied by varying the current flowing in the magnetizing winding. A third circuit is provided, having means therein for producing a variable current and means for amplifying and reproducing in the second circuit the energy variations of the current in the third circuit, the amplifying means being capable of producing materially greater energy variations than can be produced with an ordinary telephone transmitter.

1,522,286—H. P. Clausen, filed February 3, 1920, issued January 6, 1925. Assigned to Western Electric Company, Incorporated, a Corporation of New York.

METHOD AND APPARATUS FOR MOUNTING FILAMENTS in electron tubes, which consists in attaching a pair of resilient wires to the ends of the filaments and securing the wires at an angle to an intermediate portion of a pair of lead-in wires within the tube. The lead-in wires are bent to flex the resilient wires to provide tension in the filament. In this manner there is no tendency for the filament to sag over against the grid during the expansion due to heating of the cathode.

1,522,305—M. Latour, filed July 14, 1920, issued January 6, 1925.

RADIO RECEIVER, in which the received signals control the luminous condition of a thermionic filament which may be used to make an impression upon a photographic band at a radio receiving station. The incandescence of the lamp filament is controlled by the received signaling energy for setting into operation a photographic process.

1,522,308—H. F. Lowenstein and E. E. Clement, filed August 14, 1922, issued January 6, 1925. Assigned to Edward F. Colladay, Washington, D. C.

RADIO RECEIVING APPARATUS, which may be supervised in

its operation by an operator at a radio switchboard connected by line wire with the subscribers' station.

1,522,357—E. E. Clement, filed August 14, 1922, issued January 6, 1925. Assigned to Edward F. Colladay, Washington, D. C.

RADIOPHONE SYSTEM, wherein a central high power station is employed for transmission on different wave lengths. A plurality of sub-stations is provided for receiving the radio signals from the central station. Wire circuits inter-connect the central station with the sub-stations for supplying power for operation of the radio receiving sets.

1,522,358—E. E. Clement, filed August 14, 1922, issued January 6, 1925. Assigned to Edward F. Colladay, Washington, D. C.

RADIO ADVERTISING SYSTEM, which includes a combination line wire and radio communication system. The system so associates radio transmitting and receiving apparatus with a telephone trunk line that the radio operator can reach the subscriber over two paths, that is, by line wire and by space radio. The subscriber can talk to the radio operator while listening to the broadcasting, while the radio operator can connect the subscriber by line wire to the space radio system for broadcasting the conversation.

1,522,360—E. E. Clement, filed February 29, 1924, issued January 6, 1925. Assigned to Edward F. Colladay, Washington, D. C.

RADIO BROADCAST SELECTING AND DISTRIBUTING SYSTEM, in which a central station has wire connections with the subscribers' stations and supplies an unmodulated periodic current by line wire to the subscribers' stations for use as a heterodyne for incoming signaling waves. A master oscillator at the main station serves as the local source of oscillations for each of the subscribers.

1,522,361—E. E. Clement, filed February 29, 1924, issued January 6, 1925. Assigned to Edward F. Colladay, Washington, D. C.

RADIO BROADCAST SELECTING AND DISTRIBUTING SYSTEM, wherein a centralized receiving station collects energy from a distant transmission station and re-transmits the same to sub-

scribers within a local area. The power for each of the subscriber sets may be supplied from the central station.

1,522,362—E. E. Clement, filed March 22, 1924, issued January 6, 1925. Assigned to Edward F. Colladay, Washington, D. C.

SUBDIVIDED SERVICE SYSTEM OF RADIO BROADCAST DISTRIBUTION, where the central station transmits to a plurality of subscribers' stations either by currents over a line telephone circuit or space radio, the power for tha radio receiving sets being supplied from the lines. Switching means are provided for changing the tuning of the receiver when connections are changed from the antenna or line wire circuit.

1,522,581—L. Espenschied, filed August 16, 1922, issued January 13, 1925. Assigned to American Telephone and Telegraph Company, New York.

RADIO BROADCASTING SYSTEM, in which the receiving stations are enabled to have a positive indication at all times as to whether or not they are in connection with and ready to receive from the central broadcasting station. The purpose of such an arrangement is to afford a broadcasting subscriber, who may be sending out information of a business character or otherwise, a greater measure of assurance that the receiving stations to which he desires to transmit are actually receiving the information sent out. A second purpose is to give the subscriber at the receiving station definite information as to the condition of receptivity of the radio transmission channel. A break in this transmission channel may be due to some defect in the receiving station itself or some defect in the transmitting station, but in any case, it is desirable that the receiving subscriber shall be notified as to the condition of the channel. Such assurance on the part of the broadcaster is especially needed in the case of emergency alarm services as, for example, when a broadcasting station may be employed to transmit fire alarms to a number of designated receiving stations, or where police information is broadcast from a central station to a number of correlated "pick-up" points. The invention briefly consists in sending out continuously carrier waves from the broadcasting station and employing these at each of the receiving stations to operate an indicator or alarm device; in other words, to establish normally channels between the broadcasting station and all the receiving stations involved and to operate these channels on the closed

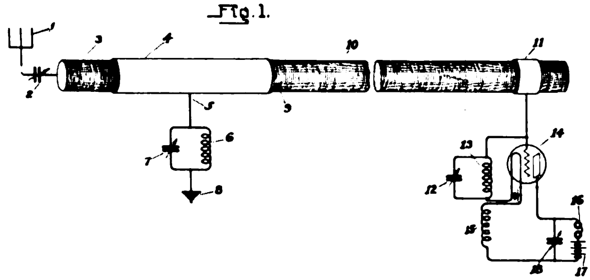
circuit basis, whereby the receiving station always definitely knows, as by the holding up of a signal, that it is in proper connection with the central station.

1,522,745—A. Press, filed June 17, 1920, issued January 13, 1925. Assigned to Westinghouse Electric and Manufacturing Company, Pennsylvania.

Balanced Antenna System, in which ground currents are substantially eliminated. The radio apparatus is connected between a grounded antenna system and circuits are provided there-between to balance out resultant current to ground.

1,522,807—L. Cohen, filed January 15, 1922, issued January 13, 1925.

Fig. 1.



Number 1,522,807—Electrical Signaling

Electrical Signaling receiving apparatus for reducing the interference arising from static disturbances. A wave-coil is employed in the receiving system. A portion of the wave-coil is enclosed in an adjustable metal tube which is electrically connected to a point on the wave-coil and grounded thru tuned circuit. The receiving apparatus is also connected in a circuit which is adjustably connected with the wave-coil. Points along the wave-coil are selected for a condition of maximum reception.

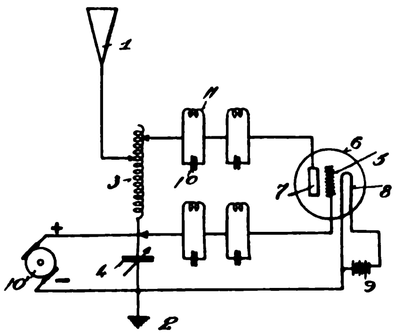
1,522,882—J. H. Hammond, Jr., filed March 16, 1912, issued January 13, 1925.

Method of and System for Selective Energy Transmission and reception, wherein the receiving station has a pair of local antenna circuits responsive respectively to high frequency undamped radiant impulses of different periodicities. A circuit is connected between the antenna circuits and is tuned to the difference between the periodicities of the received impulses for selectively observing the signals.

1,522,883—J. H. Hammond, Jr., filed June 16, 1917, issued January 13, 1925.

POLYPULSE SYSTEM OF CONTROL of the movements of a body in which a gyroscope is arranged to stabilize the body. The gyroscope is controlled by devices which respond to different frequencies radiated from a distant station whereby the direction of movement of the body may be governed.

1,523,011—W. E. Garity, filed September 23, 1920, issued January 13, 1925. Assigned to De Forest Radio Telephone and Telegraph Company, New York, a corporation of Delaware.

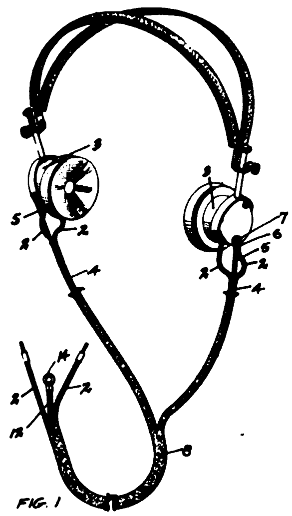


NUMBER 1,523,011—Continuous Wave Transmitting System

CONTINUOUS WAVE TRANSMITTING SYSTEM, wherein parasitic currents or harmonics of the fundamental frequency are suppressed. The transmitting system includes an electron tube operating circuit, wherein a plurality of rejector circuits are provided and each tuned to a frequency to be suppressed. The rejector circuits are connected in circuit with the grid and plate electrodes of the oscillator.

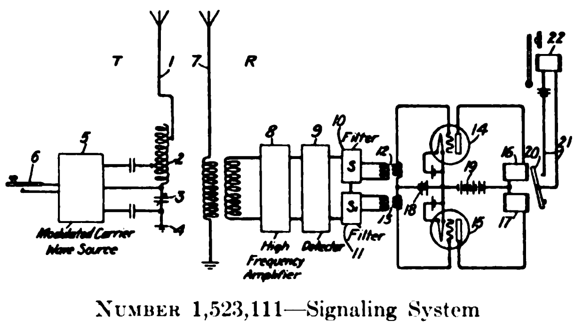
1,523,051—G. W. Carpenter and W. L. Carlson, filed June 18, 1924, issued January 13, 1925.

TELEPHONE HEADSET for use with sensitive multistage electron tube amplifiers, wherein the conductors leading to the telephone headset are shielded by a tubular webbing of a metallic tinsel and textile threads. The forming of the shield with textile threads and metallic tinsel provides an extremely flexible conductive covering for the telephone cords and permits the headset to be conveniently used in connection with the amplifier.



NUMBER 1,523,051—Telephone Headset

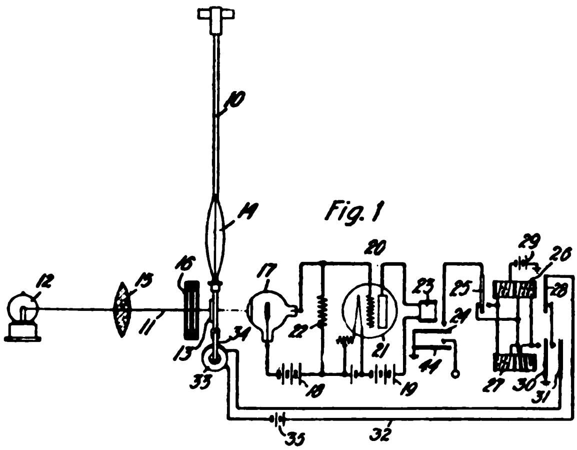
1,523,111—H. J. Fisher, filed December 5, 1923, issued January 13, 1925. Assigned to Western Electric Company, Incorporated, New York.



NUMBER 1,523,111—Signaling System

SIGNALING SYSTEM for radio telegraph operation, wherein a carrier wave is modulated by two different frequencies and the modulated and unmodulated components are transmitted to a distant station at which the corresponding components are combined to provide currents of the modulating frequencies which are selected, detected, and used to energize separate windings of a polarizing relay. The relay is employed to operate a suitable apparatus for observing the signals.

1,523,149—E. B. Wheeler, filed November 15, 1923, issued January 15, 1925. Assigned to Western Electric Company, a corporation of New York.



NUMBER 1,523,149—Means for Control of Electric Impulses

MEANS FOR CONTROL OF ELECTRIC IMPULSES in accordance with the swinging of a pendulum. The system comprises a relay for closing an electric circuit which relay may be operated by energy from a thermionic amplifier arranged to be controlled by the action of a beam of light impressed upon a photoelectric cell connected in the input circuit of the amplifier. In order to cause the relay to produce a succession of electrical impulses which shall be of equal duration and separated by equal intervals, an oscillating pendulum provided with an opaque body is arranged so that the opaque body intercepts the light beam impressed upon the photoelectric cell. This interception takes place preferably at the midpoint of the oscillation of the pendulum. It is thus evident that the light beam is twice intercepted for each oscillation of the pendulum. The relay is thereby caused to produce two pulses for each oscillation of the pendulum.

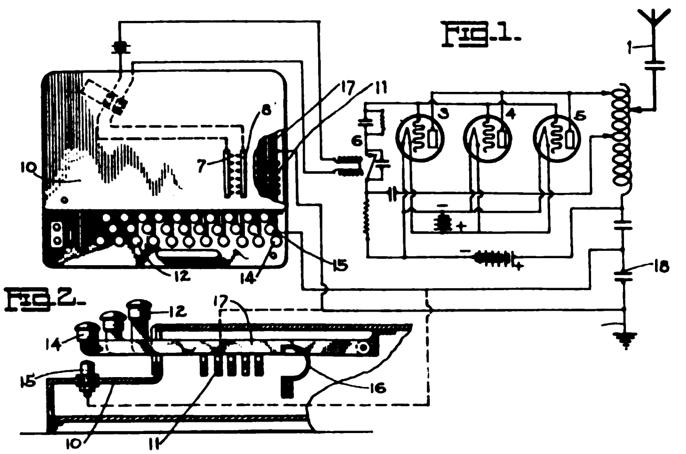
1,523,193—A. Gudheim, filed November 21, 1921, issued January 13, 1925.

AUTOMATIC FILAMENT CONTROL FOR RADIO APPARATUS, where separate rheostats are connected in each of the filament circuits and provided with a cam for operating a jack which controls the circuits to the filaments of the several tubes and insures a connection between the telephones and the last active tube in the system.

1,523,280—C. D. Palmer, filed January 29, 1921, issued January 13, 1925.

RADIO ANTENNA FOR AIRCRAFT arranged to be used either as a loop antenna or as a trailing wire antenna as may be desired. A switch is provided convenient to the operator which may be operated to connect a looped conductor on aircraft electrically as a closed circuit in the radio system or as a trailing wire balanced against the engine frame as a counterpoise for radio operation from aircraft.

1,523,377—J. B. Brady, filed August 14, 1923, issued January 13, 1925. Assigned to Morkrum Company, Chicago, Illinois, a corporation of Maine.



NUMBER 1,523,377—Radio Telegraph System

RADIO TELEGRAPH SYSTEM employing automatic printers for automatic transmission of telegraph signals from a central radio station to any number of outlying receiving stations. Means are provided for automatically starting and stopping automatic printers located at the outlying receiving stations by signals from the central station. In this manner news may be dispatched from a central radio transmitting station to the outlying printer receiving stations by preceding the printer signals with a starting signal which automatically controls the motor circuit at selected receiving stations for automatically placing the apparatus in condition for the reception of selected printer signals.

1,523,399—V. L. Chamberlin, filed February 1, 1924, issued January 20, 1925.

DETECTOR ROD SUPPORT, in which the selecting wire is carried upon a standard having universal movement for reaching any part of the crystal. The standard consists of two ball members spaced by a tubular member with the ball members connected together under tension by a coil spring. The selecting wire may be moved to a point on the crystal for obtaining maximum sensitivity.

1,523,400—V. L. Chamberlin, filed February 1, 1924, issued January 20, 1925.

CRYSTAL HOLDER for a detector which consists of a U-shaped member formed of spring metal and having inturned upper ends. A crystal support is provided within the U-shaped member and may be vertically adjustable therein for properly securing the crystal within the U-shaped member.

1,523,401—V. L. Chamberlin, filed February 1, 1924, issued January 20, 1925.

DETECTOR TIP for a crystal detector which consists of a cylindrical cap with a fine detector wire extending from the closed end thereof. A conical coiled spring extends from the end of the cap about the detector wire and fits close about the detector wire for supporting said wire intermediate its ends.

1,523,430—W. A. Knoop, filed November 20, 1923, issued January 20, 1925. Assigned to Western Electric Company, Incorporated, New York.

MOUNTING FOR VACUUM TUBES for eliminating noises in amplification systems due to shock vibration of the tube elements. The tube is provided with such a resilient support that vibrations may be damped. The tube base has a body attached thereto which projects into a pot containing a damping fluid. The body is immersed in the fluid and tends to prevent movement of the tube under mechanical vibration for suppressing noises in amplification circuits in which the tube may be connected.

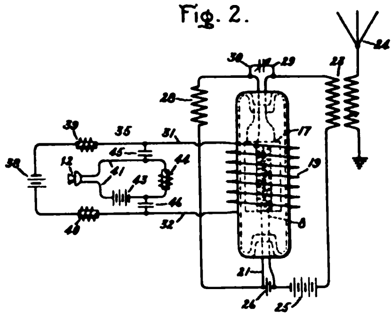
1,523,536—A. E. Greene, filed September 4, 1923, issued January 20, 1925.

ELECTRICAL CONDENSER of the variable plate variety in which the stationary plates are held in spaced relationship by means of a shearing pressure on the edge portions of the plates. A pair of supporting members is provided, forming the condenser frame

and slotted tubular members are positioned between the supporting members. Slotted members are located within the said slotted tubular members and arranged to exert a shearing pressure on the edge portions of the stator plates for maintaining the stator plates in spaced relationship.

1,523,777—A. W. Hull, filed May 21, 1920, issued January 20, 1925. Assigned to General Electric Company, New York.

Fig. 2.



NUMBER 1,523,777—Electron Discharge Device

ELECTRON DISCHARGE DEVICE, wherein an electron current is subjected to the conjoint action of a variable magnetic field. The usual tube construction is shown, including a filament, grid and plate. A grid circuit is provided for controlling an electron stream and in addition to this a winding for the development of a magnetic field is provided and arranged in the control circuit for affecting both electrostatic and electromagnetic control of the electron stream.

1,523,778—A. W. Hull, filed November 15, 1921, issued January 20, 1925. Assigned to General Electric Company, New York.

ELECTRON DEVICE AND METHOD OF OPERATING tubes of the magnetron class. The magnetron of this invention utilizes an axial magneto-strictive effect. A cathode is provided having a current-carrying capacity sufficiently high to conduct currents capable of producing a magnetic field, whereby electron current may be varied or even entirely interrupted. The cathode is surrounded by a cylindrical plate which may be connected in any desired utilization circuit.

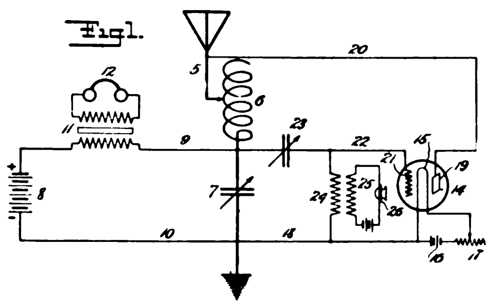
1,523,893—R. C. Pitard, filed December 15, 1923, issued January 20, 1925.

GRID CONDENSER formed by an electrical conductor adapted to embrace a portion of the tube with a dielectric material interposed between the conductor and the tube. The metal collar of the electron tube is made use of as one of the plates of the grid condenser. The other plate consists of a ribbon of metal secured around the base of the tube but separated therefrom by a dielectric sheet.

1,523,957—R. E. Hall, filed August 13, 1919, issued January 20, 1925. Assigned to Hall Research Corporation of Delaware.

TELEPHONE CALL AND METHOD THEREFOR, for summoning an operator to a receiving station by means of a signal from the transmitting station. A special form of gaseous jet relay is provided for closing a call circuit. The relay is actuated by the transmission of a particular note from the transmitting station, resulting in the tripping of the relay at the receiving station and the closing of a call circuit.

1,524,413. M. W. Sterns, filed January 28, 1920, issued January 27, 1925.



NUMBER 1,524,413—Radio Telephone System

RADIO TELEPHONE SYSTEM, in which the same tube circuit functions as a transmitter and receiver. The circuit is arranged for simultaneous transmitting and receiving. A grounded aerial circuit is provided with a tuning inductance connected thereto and a divided secondary circuit across one portion of which the input circuit of the tube is connected and across the other portion of which the output circuit is connected. A telephone modulator is inductively connected to the grid and filament circuits of the tube while the receiving device is connected in the output circuit of the tube.

1,524,645—M. Latour, filed August 19, 1921, issued January 27, 1925.

Receiving Apparatus for Eliminating Static, in which a pair of loop antennas are provided with a detector connected with each antenna and having a portion of their output circuits each common to a source of direct current. A resistance is provided in the output circuit of each detector and connections from the indicating apparatus to a point in each resistance are provided, enabling the circuits to be adjusted for the reception of signals substantially free of disturbances from static.

1,524,646—M. Latour, filed December 5, 1923, issued January 27, 1925.

Connection of Radio Frequency Alternators for Radio Signaling, in which the driving motors for the alternators are coupled together with circuits for maintaining the machines in step. A pair of alternators is shown adapted to be operated in parallel with a pair of motors connected in parallel for driving the alternators. The currents in the driving motor circuits are balanced and a supplemental circuit is provided for compensating for changes in load of the alternators for maintaining the alternators in step.

1,525,049—S. Ruben, filed June 28, 1923, issued February 3, 1925.

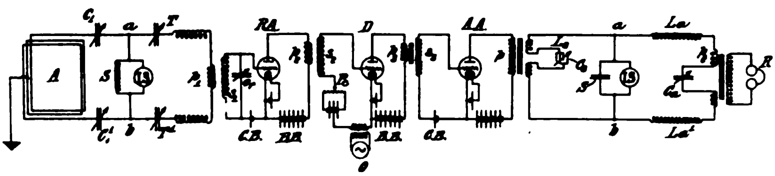
Electron Discharge Tube construction including an evacuated vessel with an electron emission element therein and a co-operating anode. An element composed of an insulating material is interposed between the electron element and the plate. A pair of electrically conductive plates are arranged in a plane approximately parallel to the anode. The electron emission element is disposed in conductive relation to the conductive plates. By this arrangement of elements the electron stream is deflected from its normal direct path and is forced to traverse an electrostatic field in a direction approximately 45 degrees to that of the field in the space between the pair of conductive plates.

1,525,110—F. K. Vreeland, filed August 6, 1919, issued February 3, 1925.

Audio Frequency Selective Signaling System, wherein the circuit connected with the detector system includes a baffle circuit and an intensity selective or other energy dissipating means which operates directly on the audio frequency currents

to exclude those currents not desired from the observing circuit. The patent sets forth the differences in characteristic between the stray impulses which are very abrupt and transitory and signaling impulses which are sustained. The strays when they constitute serious interference have an intensity or amplitude which is greater than that of the signaling impulses. The baffle circuit in the audio frequency amplifier separates the signal currents from the strays and also dissipates the strays of greater intensity than the signals, permitting the signals to be received substantially clear of static disturbances.

Fig. 3.



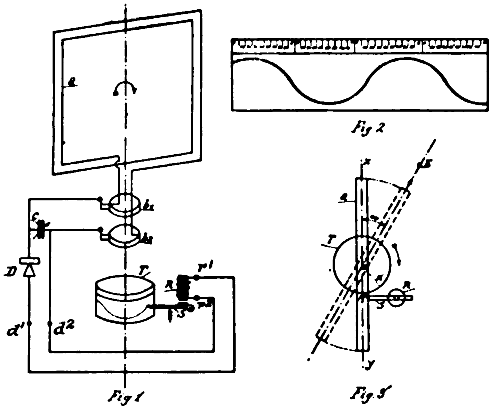
NUMBER 1,525,110—Audio Frequency Selective Signaling System

1,525,159—H. M. Wohltman and M. Hirschfeld, filed April 24, 1922, issued February 3, 1925.

CRYSTAL DETECTOR, designed to have all parts easily accessible and replaceable and in which the entire surface of the crystal may be utilized and delicate adjustment made there-against. A glass tube is provided having a large bore in one portion thereof in which the crystal is arranged to be inserted and a smaller bore in the other portion thereof thru which the contacting wire may be inserted to impress a point on the crystal. The crystal may be readily changed in position to enable the contacting wire to reach all points thereof.

1,525,177—R. B. Goldschmidt and R. Braillard, filed May 24, 1920, issued February 3, 1925.

DIRECT READING RADIOTELEGRAPHIC COMPASS, in which a loop collector is rotated about a vertical axis in order to receive maxima and minima of signal intensity. The loop frame carries a cylindrical drum on which is mounted a chart against which an inscribed pencil may be moved automatically to inscribe a curve characteristic of the received signals. In this manner the characteristics are made both visible and audible.



NUMBER 1,525,177—Direct Reading Radiotele graphic Compass

1,525,182—H. C. Hayes, filed July 7, 1923, issued February 3, 1925.

SOUND TRANSMITTER AND RECEIVER, in which a pair of annular magnets are positioned adjacent each other with a set of annular pole pieces on each of the magnets. The diaphragm consists of a magnetic annular ring which is acted upon by the annular pole pieces. The structure is designed to eliminate distortion in loud speaker reproducers.

1,525,308—A. J. Kloneck, filed November 28, 1916, issued February 10, 1925.

SIMULTANEOUS SIGNALING AND RECEIVING SYSTEM, in which a ground to antenna circuit is provided for transmitting while an antenna to antenna circuit is provided for receiving the antennas being common for both circuits. A circuit arrangement is provided between a pair of antennas for receiving signals while eliminating local disturbances from the local transmitter which may be simultaneously operating on an antenna to ground circuit.

1,525,302—R. Knopp, filed June 19, 1922, issued February 3, 1925. Assigned to Frank J. Quigan, Incorporated, Brooklyn, a firm of New York.

ELECTRIC CONDENSER of variable adjustment in which a movable plate may be shifted axially of a central shaft with respect to a stationary plate for varying the capacity therebetween. Cams are provided which are rotatable face to face for securing fine adjustment of capacity.

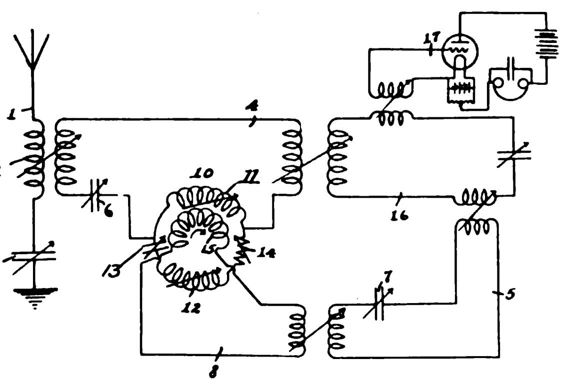
1,525,350—H. Zuckerman, filed December 4, 1920, issued February 3, 1925.

VACUUM SPARK GAP, wherein the gap is located within an exhausted tube, each gap being provided with heat radiating flanges with an insulating rod extending between the gaps for maintaining the faces in spaced relation. The electrodes are supported by U-shaped members extending from the base of the receptable.

1,525,431—T. H. Phillips, Jr., filed October 17, 1917, issued February 3, 1925. Assigned to Elmer A. Sperry, of Brooklyn, New York.

REMOTELY CONTROLLED SELECTIVE SYSTEM for radio control of circuits from a distance. The patent describes a mechanical arrangement of switches which may be operated by distinct signals from a transmitting station to close selected circuits at a receiving station.

1,525,526—J. Weinberger, filed March 3, 1921, issued February 10, 1925. Assigned to Radio Corporation of America, a corporation of Delaware.



NUMBER 1,525,526—Radio Receiving System

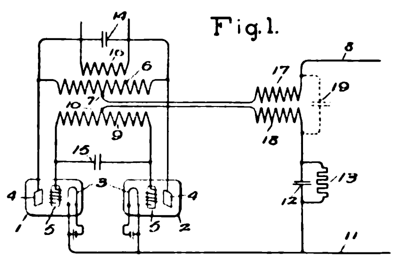
RADIO RECEIVING SYSTEM, in which relatively strong interference may be overcome while permitting the reception of desired signals. Two intermediate circuits are coupled to the receiving circuit for eliminating interfering currents from the antenna. One of the intermediate circuits has a relatively high ratio of signal current to interfering current. The other intermediate circuit is tuned to the interference and has a relatively high ratio of interfering current to signal current. The inter-

mediate circuits are arranged to oppose the intermediate currents while retaining the signal currents.

1,525,778—C. A. Hellmann, filed April 17, 1922, issued February 10, 1925.

VARIABLE CONDENSER of the rotary plate variety, wherein the two sets of elements are of unequal angular extent. Either the rotary or the stationary plates may be greater than 180 degrees in angular extent, while the other set of plates is less than 180 degrees. The object of the invention is to provide a variable condenser in which variation of capacity may be had over an angle of rotation exceeding 180 degrees and approaching as near to 360 degrees as may be desired in any particular case. The specification describes a great many modifications of the invention.

1,525,827—D. C. Prince, filed December 14, 1923, issued February 10, 1925. Assigned to General Electric Company, New York.



NUMBER 1,525,827—Production of Alternating Currents

PRODUCTION OF ALTERNATING CURRENTS by a pair of electron tubes operating in a push-pull circuit 180 degrees out of phase. A pair of three-electrode electron tubes are provided with an inductive winding having its terminals connected to the anodes of each of the tubes and an intermediate point connected to a source of direct current thru an inductance. A second inductive winding is provided which is inductively related to the first winding and has its terminals connected to the control electrodes of the tubes. An intermediate point of this last-mentioned inductive winding is connected to the cathodes thru a second inductance. The inductances are coupled to each other for the production of alternating currents by the tube system.

1,525,844—W. C. White, filed July 3, 1920, issued February 10, 1925. Assigned to General Electric Company, New York.

Vacuum Tube Apparatus, in which an electron current between two electrodes of an electron tube is controlled by means of a magnetic field. A stream of electrons is produced between a cathode and plate electrode of an electron discharge device. The velocity of this stream is caused to be greatly decreased as the electrons approach the plate electrode. By producing a magnetic field in the region adjacent to the plate electrode which acts upon these slowly moving electrons, it is possible by means of a comparatively weak magnetic field to produce a substantial variation in the electron current received by the plate electrode.

1,525,941—C. V. Logwood, filed April 21, 1920, issued February 10, 1925. Assigned to De Forest Radio Telephone and Telegraph Company, New York, a corporation of Delaware.

Radio Signaling System, in which a source of high frequency oscillations is modulated in accordance with the signals to be sent in such manner that the modulating signal current increases the power transmitted rather than decreases the same. A pair of electron tubes are provided for supplying oscillating currents to the antenna. The grid electrodes are charged with opposite potentials and condensers are provided for building up and storing energy which may be supplied to the antenna additively to the normal supply of energy thereto, whereby the modulating signal current increases the power supply.

1,526,311—M. C. Batsel, filed January 25, 1921, issued February 17, 1925.

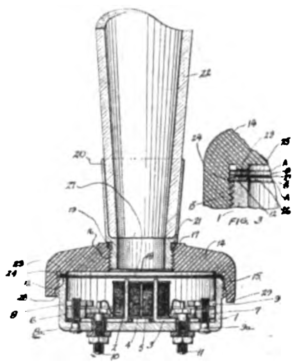
Radio Transmission System, in which a coupling system is provided between the oscillator and the radiating circuit which will keep the current in the amplifier system within safe values when the radiating system is untuned to the frequency of the oscillating circuit, but which will have the correct impedance when the two systems are tuned to the same frequency for the desired output.

1,526,408—F. W. Young, filed August 24, 1921, issued February 17, 1925. Assigned to Western Electric Company, Incorporated, a corporation of New York.

Carrier Wave Receiving System, in which de-modulators of different kinds may be used without substantial change in the receiving circuit. The invention is directed to a circuit arrange-

ment in which either a crystal detector or a tube detector may be used at will without substantial change in the receiving circuit. The tube socket is so arranged that it may receive a crystal detector support and switches are provided for adapting the circuit to a crystal rectifier instead of a tube rectifier.

1,526,626—C. E. Brigham, filed March 13, 1924, allowed February 17, 1925. Assigned to C. Brandes, Incorporated, of New York City.



NUMBER 1,526,626—Electromagnetic Sound Reproducer

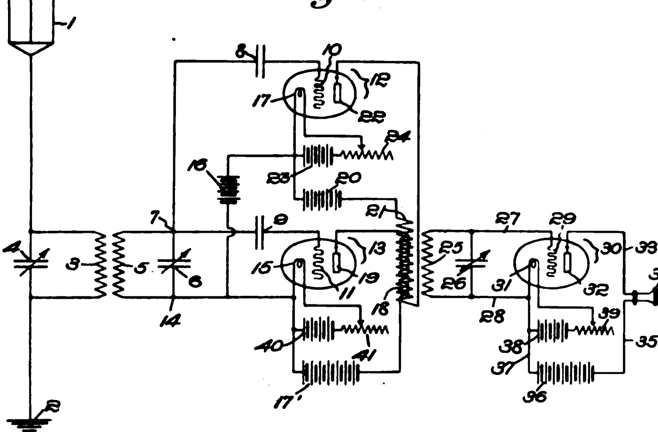
ELECTROMAGNETIC SOUND REPRODUCER for radio reception, in which a diaphragm is resiliently supported at its periphery for operation by an electromagnetic operating mechanism. The diaphragm is supported by a pair of relatively thin rings, one positioned on one side of the diaphragm and the other positioned on the opposite side of the diaphragm. The rings are composed of layers of dissimilar material formed integral with each other and remain in permanent adjustment with respect to the electromagnetic sound reproducer. This patent covers the Brandes Table Talker.

1,526,664—W. Dubilier, filed October 6, 1923, issued February 17, 1925. Assigned to Dubilier Condenser and Radio Corporation, New York, a corporation of Delaware.

ELECTRICAL CONDENSER of fixed capacity, in which the plates are clamped by the metallic sheet which wraps about the stack and forms one of the armatures of the condenser. The clamp serves as both a compression element and as a conducting plate.

1,526.852—J. H. Hammond, Jr., filed August 20, 1917, issued February 17, 1925.

Fig. 1.



NUMBER 1,526,852—Means for and Method of Limiting Interference in Radio Signaling

MEANS FOR AND METHOD OF LIMITING INTERFERENCE IN RADIO SIGNALING, by providing circuits for limiting the effect of high potentials on the detector. The receiving system includes an oscillatory circuit and a primary and secondary detector, both controlled by the oscillatory circuit. A pair of circuits are controlled by the detectors which operate upon another detector, in the output circuit of which the observing instrument is connected. The detectors are arranged to be responsive in different degrees to impulses having a given intensity.

1,527,228—J. C. Schelleng, filed December 29, 1923, issued February 24, 1915. Assigned to Western Electric Company, Incorporated, New York.

METHOD OF HARMONIC OR SUB-HARMONIC FREQUENCY PRODUCTION by a circuit arrangement utilizing an oscillator for producing a complex wave the fundamental frequency of which is normally approximately that of the desired sub-harmonic frequency and having a harmonic approximately equal to the reference frequency. The current of the reference frequency is caused to coact with the current of the harmonic frequency in the oscillation circuit of the oscillator. The effect of such coaction on the oscillator is to pull it into step with the frequency of the impressed current, that is, to cause the reference and harmonic frequencies

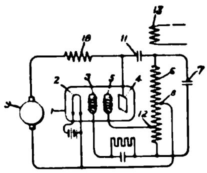
to become identical. The fundamental frequency is thus correspondingly automatically adjusted to exact sub-harmonic relationship with the reference frequency. The production of the unmodulated carrier may be accomplished by means of the present invention by combining the received side bands, selecting the resulting component of double carrier frequency and causing it to operate upon a local oscillator in the manner described above to produce the first even sub-harmonic, that is, the desired carrier frequency.

1,527,578—D. H. Sheriff, Jr., filed March 5, 1923, issued February 24, 1925.

Variable Condenser, designed for the securing of fine or critical adjustment after the usual rotative adjustment has been made. The plates are formed so that one set may be rotated with respect to the other set and then one set may be moved out of and into parallelism with the other set of plates to effect fine adjustment of capacity.

1,527,703—D. C. Prince, filed April 8, 1922, issued February 24, 1925. Assigned to General Electric Company, New York.

Fig. 1.



Number 1,527,703—Electron Discharge Apparatus

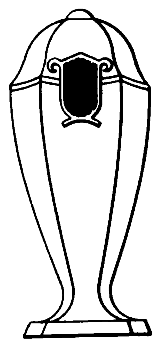
Electron Discharge Apparatus, in which losses incident to the operation of electron tube discharge devices are eliminated. Losses which occur by reason of secondary emission are prevented by interposing a supplemental grid between the control electrode and the plate. If, when current is flowing to the plate, this extra grid is made slightly negative with respect to the plate no secondary electrons can escape from the plate and the control grid may be made positive enough greatly to reduce the space charge losses without introducing compensating grid losses.

1,527,896—S. L. Miller, filed June 7, 1922, issued February 24, 1925.

Radio Cabinet, designed to facilitate the quick and easy

positioning therein of the radio apparatus. The front side of the cabinet is open. A panel is provided having a sub-base at its lower portions slidable in the base of the cabinet. The cabinet is provided with a plurality of openings in its rear side through which binding posts carried by the slidable sub-base may be projected for completing connections with the apparatus from the exterior of the cabinet.

Design 66,668—Stephen Bourne, filed November 26, 1924, allowed February 24, 1925. Assigned to C. Brandes, Incorporated, of New York City.



NUMBER 66,668—
Design for a Loud
Speaker for Radio
Reproduction

DESIGN FOR A LOUD SPEAKER FOR RADIO REPRODUCTION, in which the sound is directed upwardly in a vase-like amplifying horn and outwardly to the listener-in.

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 13 JUNE, 1925 Number 3

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LONG DISTANCE RADIO RECEIVING MEASUREMENTS IN 1924*

By

L. W. AUSTIN

(LABORATORY FOR SPECIAL RADIO TRANSMISSION RESEARCH, WASHINGTON D. C.)

(Conducted jointly by the Bureau of Standards and the American Section of the International Union for Scientific Radio Telegraphy.)

Two stations, Monte Grande (LPZ), Argentina, and Cayey (NAU), Porto Rico, have been added during the year to the number of those regularly measured in Washington. Monte Grande is interesting; first, because it is 2,000 km. farther away than the European stations, and second, because the waves travel in a south-north direction from the southern to the northern hemisphere; thus encountering entirely different seasonal conditions from those encountered in the transmission from Europe to America. The station gives nearly the same morning intensity as Nauen, Germany, and the ratio of average observed to calculated values is about three to one. Unfortunately, Monte Grande does not send in the afternoon.

Cayey has been observed partly because its frequency, approximately 33.8 kc. (8,870 m.), is considerably higher than the other stations and partly on account of its nearly south-north direction of transmission, which at certain seasons lies nearly parallel to the sunset shadow wall. It was thought that this might cause eccentricities in reception at about sunset, but no peculiarities have been observed on the rather limited number of occasions when transmission took place at that time.

The mean monthly values of the field intensities of the signals from the various stations, and of the corresponding atmospheric disturbances, are shown in the tables and curves.

Table I gives the approximate data concerning the transmitting stations, as far as known.

*Received by the Editor, February 20, 1925. Published by permission of the Director of the Bureau of Standards of the United States Department of Commerce.

TABLE I

APPROXIMATE TRANSMISSION DATA

	Frequency f kc.	Wave Length λ m.	Antenna Current I amp.	Effective Height h m.	Distance d km.
Nauen POZ[1]	23.4	12,800	390	145	6,650
Bolinas KET	22.9	13,100	420	51	3,920
Cayey NAU	33.8	8,870	150	120	2,490
Monte Grande LPZ	23.6	12,700	610	150	8,300
Lafayette LY	15.9	18,900	475	180	6,160
Ste. Assise UFT	20.8	14,400	380	180	6,200
Ste. Assise UFU	15.0	20,000	475	180	6,200
Malabar PKX	19.0	15,800	500	320	14,700
Cavite NPO	19.3	15,500	180	120	11,800

Tables II and III give the monthly averages of the received field intensities and of the corresponding atmospheric disturbances in microvolts per meter. It is to be remembered that the signals received in Washington at 10 A. M. from Europe have an all-daylight path, tho during the short days of winter they are probably disturbed by being transmitted too close to the European sunset time. The 3 P. M. signals are sent during the evening hours and during the winter considerable parts of their paths lie in darkness.

TABLE II

AVERAGE SIGNAL AND ATMOSPHERIC DISTURBANCE INTENSITIES IN 1924 FOR LAFAYETTE (LY), STE. ASSISE (UFU), AND EL CAYEY (NAU) IN MICROVOLTS PER METER

1924	A. M.			P. M.			A. M.		P. M.	
	LY	UFU	Dist.	LY	UFU	Dist.	NAU	Dist.	NAU	Dist.
January	130.0	63.5	21.2	160.0	89.6	26.5	....	....	....	....
February	153.0	64.2	39.3	125.7	71.5	70.6	....	....	....	....
March	117.5	50.3	30.2	88.3	46.7	70.4	....	....	....	....
April	136.7	50.9	65.8	88.2	34.7	166.5	73.2	17.7	59.3	47.3
May	107.5	52.2	97.3	75.8	34.7	180.0	79.8	23.3	59.9	64.4
June	120.0	45.8	105.4	77.3	36.6	605.0	57.3	35.4	43.5	170.0
July	113.6	47.1	56.0	61.8	22.5	267.0	112.5	30.0	66.5	137.0
August	93.5	40.3	87.0	52.5	17.7	294.0	57.0	42.0	73.2	157.0
September	119.7	55.3	50.0	88.6	35.3	151.0	100.2	19.0	92.5	88.0
October	113.7	54.4	46.0	137.4	57.0	110.0	87.0	10.0	67.6	31.0
November	87.8	37.4	38.3	180.9	66.3	66.0	62.8	10.8	65.1	14.0
December	87.6	50.3	30.2	151.5	64.2	35.9	56.1	7.1	62.1	7.8
Average	115.0	50.9	55.5	107.3	48.0	170.2	76.2	21.7	65.5	79.6

Figure 1 shows the monthly averages of the 10 A. M. signals from Bordeaux for the years 1922, 1923, and 1924. Figure 2 gives similar 10 A. M. curves for Nauen. Figure 3 shows the

[1] During the year Nauen has used at times an antenna with $h = 175$ m. and a current varying between 300 and 480 amperes for its 23.4 kc. frequency.

TABLE III

AVERAGE SIGNAL AND ATMOSPHERIC DISTURBANCE INTENSITIES IN 1924 FOR STE. ASSISE (UFT), BOLINAS (KET), NAUEN (POZ), AND MONTE GRANDE (LPZ) IN MICROVOLTS PER METER

1924	A. M.					P. M.				
	UFT	KET	POZ	LPZ	Dist.	UFT	KET	POZ	LPZ	Dist.
January	37.3	....	16.5	....	17.7	39.3	....	24.0	....	22.1
February	41.8	80.4	19.3	....	32.8	37.4	69.6	21.8	....	61.3
March	40.7	59.9	35.7	40.1	24.0	32.5	54.4	37.0	....	58.9
April	39.6	56.3	28.5	33.6	53.3	21.2	47.4	17.8	....	136.0
May	34.3	57.4	22.5	27.2	77.3	21.4	40.6	13.1	....	158.0
June	37.0	55.4	27.1	26.1	90.4	21.5	35.0	18.9	....	531.0
July	43.1	53.5	30.3	33.3	50.0	22.6	26.1	15.1	....	238.0
August	40.4	24.5	34.9	41.3	78.0	21.8	36.3	15.7	....	306.0
September	59.5	58.9	49.6	39.3	54.0	36.5	50.8	31.2	....	148.0
October	49.9	62.7	31.2	42.3	31.0	44.0	59.8	32.0	....	81.0
November	24.3	49.3	14.5	38.4	26.0	39.0	61.2	43.8	....	56.0
December	32.7	54.2	21.7	46.7	23.5	41.4	54.1	39.4	....	25.9
Average	40.0	55.6	27.6	36.8	46.5	31.5	48.6	25.8	....	156.0

morning signals for Cayey and Monte Grande. Figure 4 shows the variations in the monthly disturbance averages at 3. P. M. for three frequencies. The marked difference between the disturbances at 24.0 kc. and 33.3 kc. is noticeable. In Figure 5 the 3 P. M. disturbances for a frequency of 24.0 kc. (12,500 m.), are plotted for the years 1922, 1923, and 1924. Figure 6 gives similar curves for a frequency of 15.0 kc. (20,000 m.), for the same years

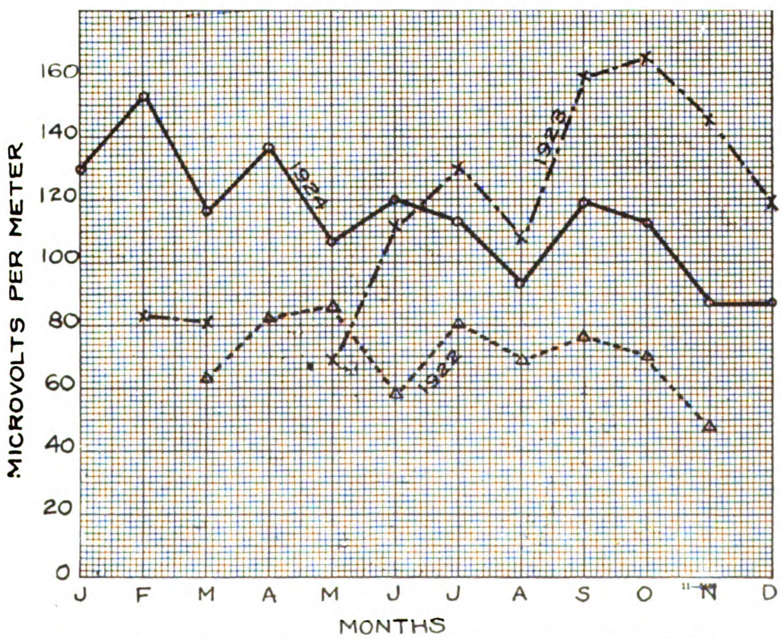

FIGURE 1—Lafayette (LY) Average Signal 10 A.M. 1922–1923–1924

These curves for 1922, 1923, and 1924, are given merely as information. It is too early to attempt to draw any definite conclusions from their variations.

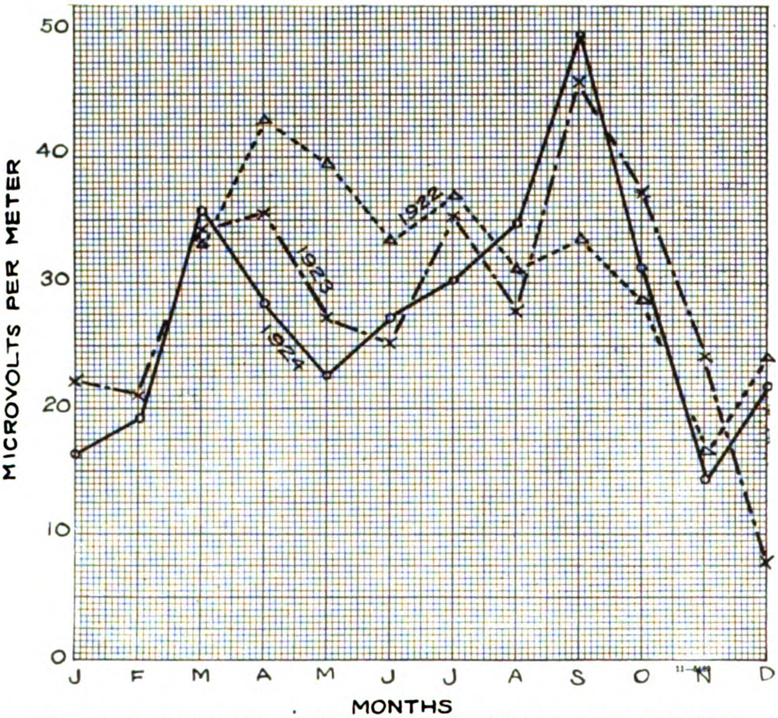

FIGURE 2—Average Signal for Nauen (POZ) 10 A.M., 1922–1923–1924

Field intensity measurements were made during August and September at San Diego, California, on the high-power arc stations, Cavite, Philippine Islands, and Malabar, Java. The distance from Cavite to San Diego is approximately 11,800 km. (6,400 nautical miles), with a difference in time of eight hours, while the distance from Malabar is 14,700 km. (8,000 miles), with a difference in time of nine hours. This is about the greatest distance which can be attained for all daylight and approximately all water communication with the present high-power stations of the world. Even in this case there are only about two hours during the day available for observation without too close approach to sunset or sunrise at one station or the other. The observations were taken with the telephone comparator and the apparatus calibrated with a radio-frequency generator and attenuation box as in the Western Electric method of measuring signals.

The final results were as follows:

	Cavite	Malabar
Observed averages..................	2.04 $\mu v./m.$	4.02 $\mu v./m.$
Calculated (Austin-Cohen formula)..	0.69	1.83

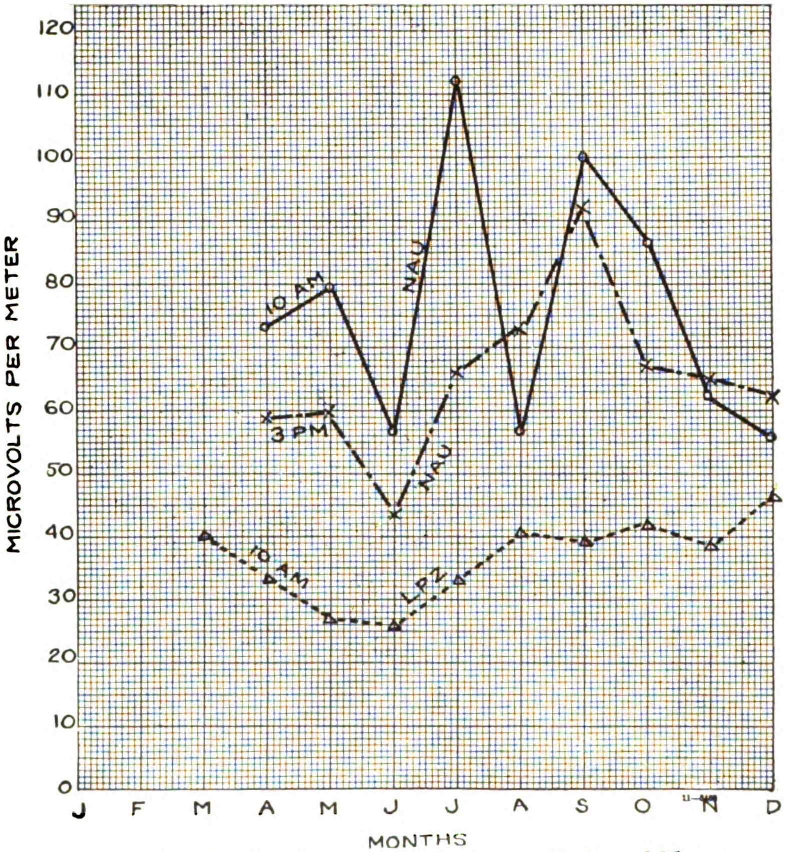

FIGURE 3—1924 Signal Averages of El Cayey (NAU) and Monte Grande (LPZ)

During the year experiments have been carried on to determine the effect of heavy atmospheric disturbances on the observed values of the strength of signals by making measurements on the telephone comparator, first with an artificial antenna and then with an elevated antenna on which the disturbances were

coming in. It was found: (1) That if the disturbances were separated by intervals of comparative silence, the readings were independent of the intensity of the disturbances provided the telephones were removed from the ears sufficiently to prevent the deafening effect of the crashes. (2) If the disturbances were

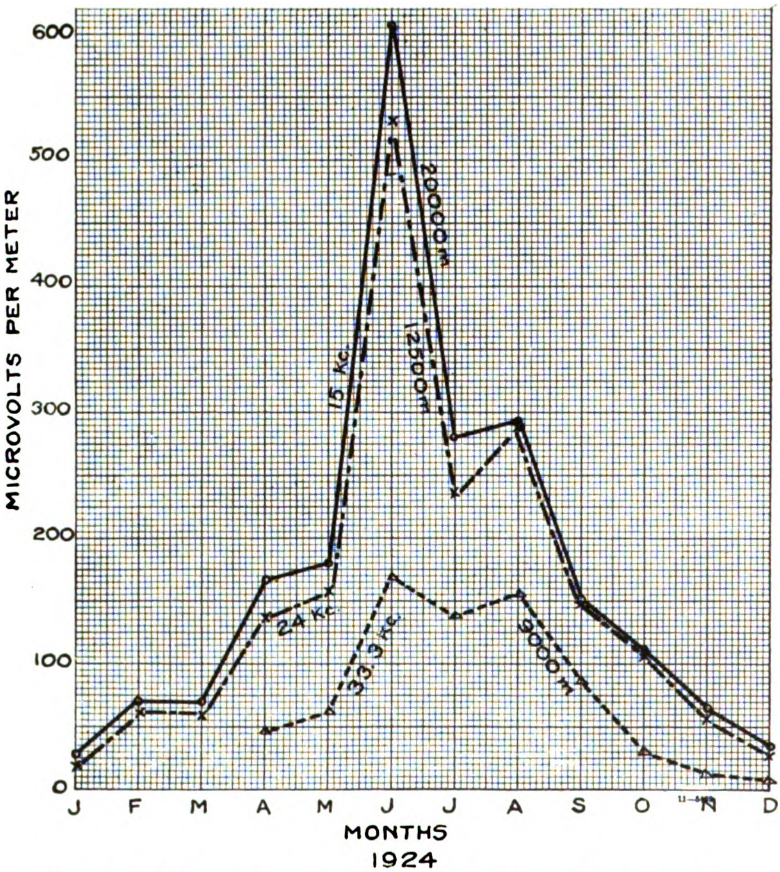

FIGURE 4—1924 Average Atmospheric Disturbances 3 P.M. for 15 kc. (20,000 m), 24 kc. (12,500 m), and 33.3 kc. (9,000 m)

practically continuous but less than about seven times the strength of the signal, the observations were unaffected. (3) With the continuous disturbances between seven and sixteen times the strength of the signal the observed values are too low. (4) When the disturbances are more than sixteen times the signal strength, the signal is not heard. These experiments have made it possible to make estimates of the signal strength of the weaker

stations on the summer afternoons instead of arbitrarily throwing them out, or considering them inaudible. This is a matter of some importance for the determination of the summer afternoon fading. The application of these corrections to the afternoon observations of 1922 and 1923, practically doubles the average values of the summer afternoon readings of the weaker stations, like Nauen.

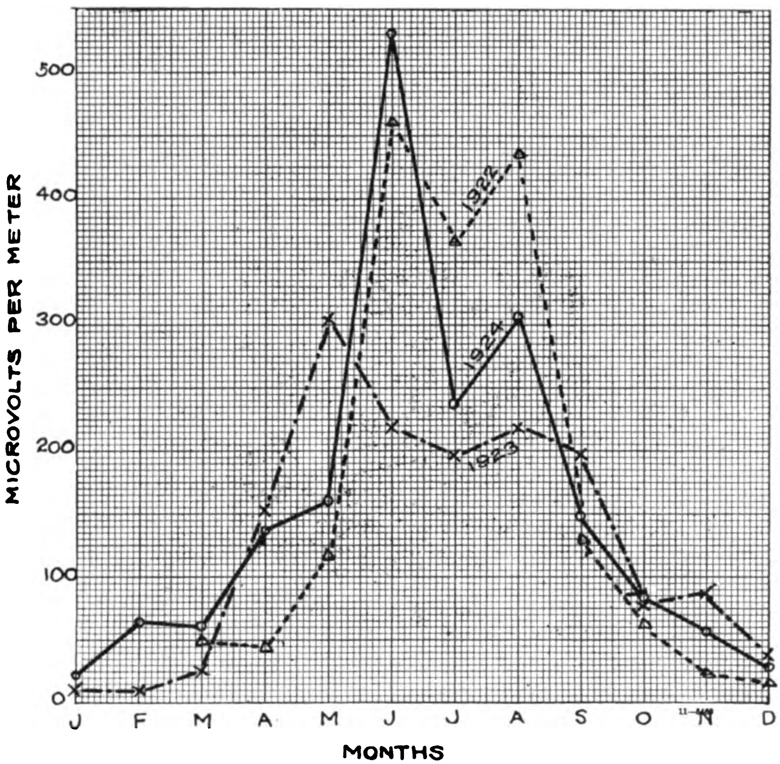

FIGURE 5—Average Atmospheric Disturbances 3 P.M. for 1922–1923–1924
f = 24 kc. (12,500 m)

Some work has been done during the year on the weakening of the European stations at about the time of the European sunset. This plays a part in the production of the weak signals observed at 10 A. M., in November, December, and January, and in the afternoon fading observed on the 3 P. M. signals in summer. On account of the limited personnel of the laboratory, it has not been possible to complete this part of the work for presentation.

The similarity in the monthly average intensity curves taken

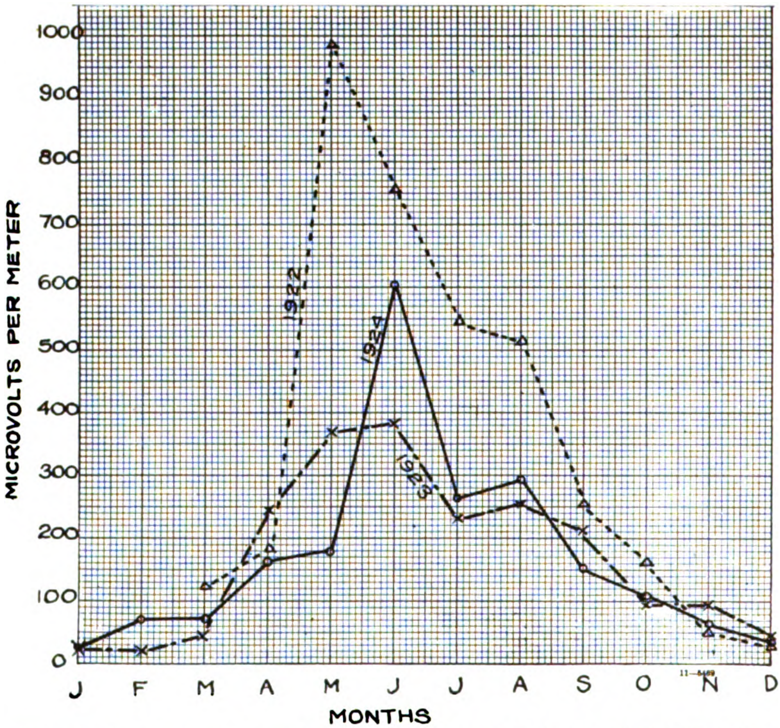

FIGURE 6—Average Atmospheric Disturbances 3 P.M. for 1922–1923–1924
$f = 15$ kc. (20,000)

at Meudon and in Washington on the U. R. S. I. signals sent out from Bordeaux (LY) at 3 P. M., Washington time, has continued to be worthy of note. This similarity began to be observed at the time of Bordeaux's change in frequency from 12.8 kc. (23,400 m.), to 15.9 kc. (18,900 m.), in May, 1923, as was mentioned in last year's report. Similar, nearly simultaneous readings have also been taken on Rocky Point, L. I. (WQL) at Meudon and Washington, but in this case no definite correspondence between the two reception curves has been found.

Bureau of Standards,
January, 1925.

SUMMARY: The paper gives a resumé of the measurements of 1924. Observations were taken twice a day on the following stations: Lafayette (Bordeaux), Ste. Assise (Paris), Nauen (Berlin), Bolinas (California), Monte Grande (Argentina), and Cayey (Porto Rico). In addition, measurements were made in August and September at San Diego, California, on signals from Cavite, P. I., and Malabar, Java, respectively, 11,300 and 14,700 km. distant. The tables and curves show the monthly averages of signal and corresponding atmospheric disturbance strength for the various stations and also some comparisons of these quantities for the years 1922, 1923, and 1924.

PRODUCTION OF SINGLE SIDEBAND FOR TRANS-ATLANTIC RADIO TELEPHONY*

By

R. A. Heising

(Western Electric Company, New York)

On January 5, 1923 the first public demonstration was made of the use of the single sideband eliminated carrier method of transmission applied to radio. The occasion of these tests was the transmitting to England of messages spoken by the officials of the American Telephone and Telegraph Company from their offices at 195 Broadway, New York. These tests, which were made possible by the co-operation between the engineers of the American Telephone and Telegraph Company and the Western Electric Company and the engineers of the Radio Corporation of America, have been described in a paper by Arnold and Espenschied.[1]

This single sideband eliminated carrier method of transmission has been in use on wires for several years. This method was invented by J. R. Carson. It is described in his patent[2] and is discussed in a paper by Colpitts and Blackwell.[3] The electrical filter which plays an important part in the system here described was invented by G. A. Campbell. Its advantages over the ordinary modulated carrier system of radio transmission are such as peculiarly to fit it for long wave radio telephone work.

The Single Sideband System

The general principles of the system, whether applied to wire or radio communication, are outlined in the papers just referred to and in one by Hartley, but as the application which I am about to describe is best understood by having a simple point of view, it is desirable to repeat a certain amount of what has been given in these papers.

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*Presented before The Institute of Radio Engineers, New York, March 19, 1924 Received by the Editor, January 10, 1925

[1] "Journal of the American Institute of Electrical Engineers," June, 1923.
[2] United States Patent Number 1,449,382, also 1,343,306 and 1,343,307.
[3] "Journal of the American Institute of Electrical Engineers," April, 1921.

To begin with, we must consider the nature of our signals to be transmitted. The sound waves in speech are exceedingly complex. It has been found, however, that with a continuous band of frequencies extending from 200 cycles to 2500 cycles intelligible transmission of speech can be obtained. While entirely perfect reproduction of speech would require an extension in the frequency band in each direction, the range given above is sufficient for reasonably satisfactory communication. In reducing the construction of speech to this sustained wave basis we can use a method of handling the analysis of our system which is relatively easy in the present state of the art.

When several people speak simultaneously, frequencies occurring in their various voices all fall within the same range. Any system of communication built to transmit several conversations over the same wire or thru the same medium must provide means for sending several of these groups of frequencies without their mutually interfering at the receiving end. That is, if the frequency range representing conversation *A* in Figure 1 is a group occurring in one conversation, we must make it possible to keep this group separate from other groups that fall in the same range. What we call the single-sideband system provides a method for doing this as follows: Suppose we take the group *A* and shift it abruptly to the position marked *B* which is, say 25,000 to 30,000 cycles. A second conversation also falling in the region *A* is then shifted by another piece of apparatus up to the position *C* which runs from 20,000 to 25,000, and so on in like manner other conversations can be given positions *E*, *D*, and so on. Now with this shift in the frequencies to higher values, each conversation occupies its own frequency region and it is possible by the use of filters to separate one group from another. That is, at a receiving end group *B* will be selected by a suitable filter which discriminates against *E*, *C*, and so on, and when this group is shifted back to its original position it is understandable to the recipient. Simultaneously *C* and *E* selected by their respective stations are also moved back to the positions originally held at *A* and are also understood by their respective recipients. There is then no mutual interference between these various conversations.

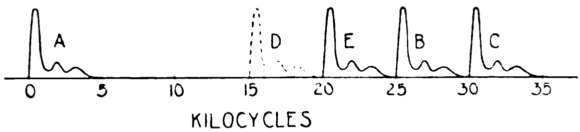


FIGURE 1—Shifted Speech Frequency Bands for Multiplex Communication

This system of communication has certain advantages over ordinary radio telephony. One is that the frequency band region which it occupies is only equivalent to that taken up by speech itself. It is one-half of the frequency band width which is taken up by the ordinary modulated carrier in radio, so that there is a doubling of the number of channels of conversation over what ordinary radio will allow. Another advantage is that there is no carrier transmitted. In the ordinary modulated wave at least two-thirds of the energy goes into the carrier which is not one of the intelligence-carrying frequencies. In this new system all of the energy goes into those frequencies which represent the speech frequencies and when converted back into speech frequencies at the receiving end provide those frequencies with the necessary energy. In view of the fact that the power used for trans-Atlantic communication runs up to the order of hundreds of kilowatts, it is highly desirable in the interest of economy that as little be put into the non-intelligence-carrying frequencies and as much be put in intelligence-carrying frequencies as is possible.

A third advantage is that with the narrowing of frequency range which occurs over what we get in ordinary modulated radio it is possible to work on sharper tuned antennas at long waves. This is of particular importance.

Method of Production

The sliding of a conversation band such as represented by frequency group A up to range B is accomplished by making use of well-known principles. When a sustained wave has its amplitude varied in accordance with an audio frequency signaling wave the resulting modulating wave represented by equation

$$i = A \sin \omega t (1 + \Sigma K \cos \phi t)$$

may be looked upon as a group of frequencies of steady amplitude. The principal frequency is the carrier frequency, $A \sin \omega t$, and on either side of this carrier, groups of frequencies known as side-bands occur. A sideband group of frequencies consists of an aggregation of frequencies having exactly the same frequency-amplitude distribution when measured from the carrier frequency position as has the speech signal $\Sigma k \cos \phi t$. That is, if conversation A modulates a carrier C in Figure 2 there are produced above frequency "C" a group of frequencies called the upper sideband having frequencies $\left(C + \frac{\phi}{2\pi}\right)$ and below frequency C another group called the lower sideband, having frequencies $\left(C - \frac{\phi}{2\pi}\right)$

These frequencies above and below the carrier occur simultaneously with the frequencies in the speech group and have relatively the same amplitude and variation in amplitude with the frequency. We, therefore, usually say the frequencies produced when a radio frequency wave is modulated are the carrier frequency, the carrier plus the speech frequencies, and the carrier minus the speech frequencies.

The ordinary process of modulating a carrier wave thus produces a group of frequencies such as *B* (Figure 2) having exactly the same frequency-amplitude arrangement as the original speech frequencies, but they occur in a totally different part of the frequency range. This gives us the shift in the speech band which we desire. The undesirable things about it, however, are that they are located close by a carrier frequency and the frequencies of the other sideband. To get the desired band isolated, it is necessary to discriminate against or eliminate entirely the carrier and the undesired sideband.

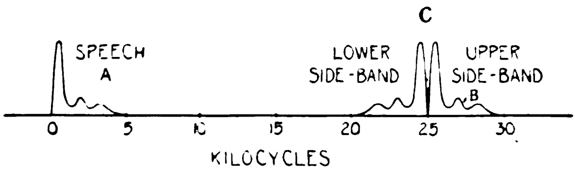


FIGURE 2—Sidebands Produced When a Radio Frequency Carrier Wave is Modulated by speech

The elementary system of a single sideband radio telephony is, therefore, to modulate a carrier with speech, put the modulated wave thru a filter which will eliminate the carrier and undesired band and then amplify the desired band and put it into the antenna. Practically, it is not so simple. In its application to line wires the carrier is first eliminated by balanced tube circuits. In the application of it to radio further complications become necessary.

The difficulty is apparent when we consider that the upper and lower sidebands lie quite close together. If we wish to use a sideband near 50 000 or 60,000 cycles we would find a satisfactory filter prohibitively costly to build in the present state of the art. Also, if we rely on a single filter to accomplish our selection, either the filter must be adjustable or else transmission must be restricted to one frequency. It would be an obvious hardship in radio not to be able to change wave length to avoid interference, and on the other hand adjustable filters are difficult and

costly to construct. We, therefore, resorted to the process of double modulating[5] in addition to using a balanced modulator for carrier elimination in order to get around both the difficulties mentioned above.

Carrier Elimination

To eliminate the carrier we can use several circuit schemes. For instance, in Figure 3 is shown a bridge arrangement in the

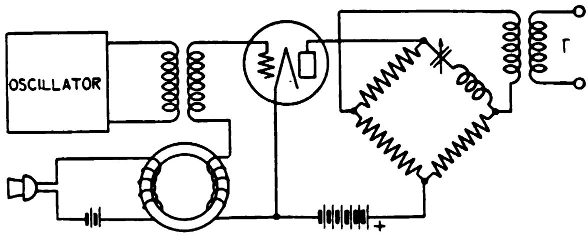


Figure 3—Modulator Tube with Bridge Circuit to Eliminate the Carrier

plate circuit of a modulator. The bridge contains one arm tuned to the carrier, and is balanced for it. The carrier will not be present in the output of this bridge network or on the grid of the amplifier. For the sideband frequencies, however, the bridge will be unbalanced and they will be impressed by means of the transformer, *T*, on the amplifier. Or we might use the circuit shown in Figure 4, where the output of the modulator tube *M* is

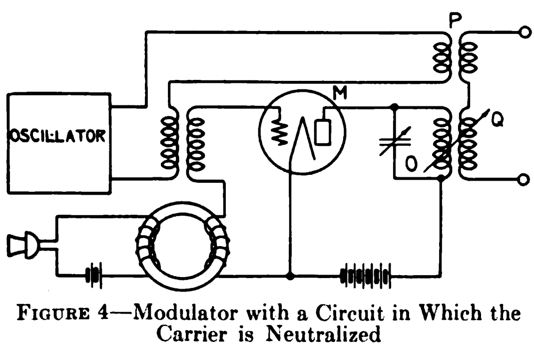
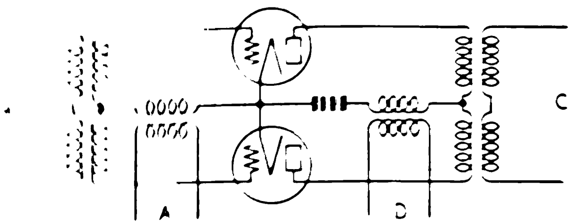


Figure 4—Modulator with a Circuit in Which the Carrier is Neutralized

impressed on the circuit "*Q*." A coil *P* connected to the oscillator couples with an inductance in circuit "*Q*" and introduces enough of the carrier in opposite phase to neutralize the carrier from the tuned circuit "*O*." In the case of the side frequncies produced

[5] Espenschied, United States Patent Number 1,447,204.

[illegible] tube and delivered through the circuit "O." [illegible] to balance them out and they will be im- [illegible] amplifier. Or we might make use of the bal- [illegible] shown in Figure 5. In this case two modulator [illegible] The carrier is put on from the oscillator to the [illegible] while the speech comes in on the grids in [illegible] transformer C in the output circuit will there- [illegible] any of the carrier but it will transmit the side- [illegible] other modifications of these circuits which [illegible] about the same results, but these are the principal [illegible] in Figure 5 is the one we have used in our work.



[illegible] Balanced Modulator Showing the Two Places [illegible] the Speech and Carrier, and Two Places for securing the Output

[illegible] to which the carrier must be eliminated is not so [illegible] the undesired sideband. It gives trouble only in pro- [illegible] its magnitude relation to the locally supplied carrier [illegible] receiving end. That is, the amplitude of any audio fre- [illegible] beat note it produces is

$$A = C\,S\,k$$

[illegible] is the carrier amplitude received and S is the sideband amplitude received. The locally supplied carrier C' will produce [illegible] beat note of

$$A' = C'\,S\,k$$

[illegible] always is much greater. The trouble comes only if [illegible] and C' do not have identical frequency values. If the ordinary detector is used in receiving, it is necessary to make C as small as possible so that C' does not have to be exactly in synchronism. If we use a balanced detector the magnitude of C is of secondary importance as the circuit will automatically eliminate A but not [illegible]

In keeping our carrier down to a minimum we make use of both the balanced modulator and of the filter. The balanced modulator reduces the carrier to a very considerable extent and

by placing the carrier about half way up the side of the filter attenuation curve, it is reduced a considerable amount more. The final value is then a fraction of a percent of the value it would have if not reduced at all. In order to eliminate the carrier sufficiently a proper proportion of the carrier and speech amplitude must be made. The magnitude of the sideband is proportional to the product of the carrier and signal amplitudes, while of course any unbalance and frequency selection in the carrier elimination circuit will give a carrier output proportional to the carrier input, so that we do not want to make the carrier too large, because any unbalance will allow a proportionately large amount to get thru.

THE BALANCED MODULATOR

A little further information on balanced modulators will probably not be out of place at this point. Suppose we take a circuit as represented in Figure 5 and add a transformer in the position *D*. We then have two places in which to impress voltages on the grids *A* and *B* and two places to take the power out of the circuit, *C* and *D*.

If there are impressed a carrier voltage E_1 and a speech voltage E_2 we write the equation of current as:

$$
\begin{aligned}
i &= a_1 (E_1 \cos \omega t + E_2 \sin \phi t) + a_2 (E_1 \cos \omega t + E_2 \sin \phi t)^2 + \dots \\
&= a_1 E_1 \cos \omega t + a_1 E_2 \sin \omega t + a_2 E_1^2 \cos^2 \omega t \\
&\quad + a_2 E_2^2 \sin^2 \phi t + 2a_2 E_1 E_2 \sin \phi t \cos \omega t + \dots\dots \\
&= a_1 E_1 \cos \omega t + a_1 E_2 \sin \phi t \\
&\quad + \frac{a_2 E_1^2}{2} + \frac{a_2 E_1^2}{2} \cos 2 \omega t \\
&\quad + \frac{a_2 E_1^2}{2} - \frac{a_2 E_2^2}{2} \cos 2 \phi t \\
&\quad + a_2 E_1 E_2 \sin (\omega + \phi) t \\
&\qquad a_2 E_1 E_2 \sin (\omega + \phi) t + \dots\dots
\end{aligned}
$$

for tube number 1. If we had both the carrier and the speech frequencies impressed on transformer *B*, the current in tube number 2 would be identical except that the signs of E_1 and E_2 would be reversed. The current taken out thru transformer *C* would be the difference of these two currents, while the current taken out thru transformer *D* would be the sum of the two. If, however, we both put the speech and carrier in at transformer *A*, the signs of E_1 and E_2 in the equations for the currents will be the same, and the sum and difference currents coming out thru transformers *D* and *C* will be quite different. There are thus totally

different currents to be secured at these two outputs depending upon where the inputs occur. Also E_1 and E_2 may be put in at separate transformers *A* and *B*, respectively or vice versa, and in that case still different results occur. The four combinations possible are shown in table *A*.

TABLE A

	Speech in at	Carrier in at	Out at *C*	Out at *D*
1	*A*	*A*	0	ϕ, ω, 2ϕ, 2ω, $\omega-\phi$ $\omega+\phi$
2	*A*	*B*	ω, $\omega+\phi$, $\omega-\phi$	ω, 2ϕ, 2ω
3	*B*	*A*	ϕ, $\omega+\phi$, $\omega-\phi$	ω, 2ϕ, 2ω
4	*B*	*B*	ϕ, ω	2ϕ, 2ω, $(\omega-\phi)$, $(\omega+\phi)$

All odd harmonics of ω and ϕ come out with these respective frequencies, and the even harmonics come out where 2ω and 2ϕ, respectively, are shown.

If our only purpose is to eliminate the carrier ω, $2n$ we can use either combination 3 or 4 from the table. We would take the sidebands out thru transformer *D* if using combination 4. However, the second and even harmonics come out with the sidebands in combination 4 while only the speech frequencies (and odd harmonics if present) come out with the sidebands in combination 3. As the transformer to handle the side frequencies will be inefficient for the speech frequencies, we can secure the sidebands free of other frequencies more easily with combination 3 than with any of the others.

Filtering

The principal reason why we do not use the more simple process of producing the single sideband is that it is too expensive to build filters sufficiently sharp to separate one sideband from another at carrier frequencies up in the neighborhood of 60,000 cycles. In order to get a single sideband at 60,000 we resort to the process of modulating twice. That is, we secure our single sideband at a low enough frequency to separate it easily from the carrier and the other sideband and then by a second modulation process we move it to the desired point. This is represented in Figure 6. The speech band represented by *A* is used first to

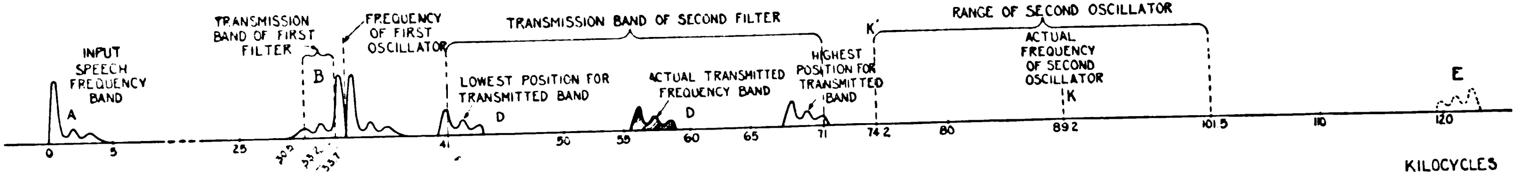


FIGURE 6—Positions of the Various Sidebands and Carriers in the Double Modulation Process

modulate a carrier such as 33,700 cycles. There are then produced an upper and a lower sideband at that frequency. It is comparatively easy to separate the bands at this frequency. In this particular case we pick out the lower sideband, that is, we use a filter which transmits the frequencies running from 30,500 to 33,200. For this purpose the filter is built with a good steep slope on the upper side. The filter which we use has an attenuation characteristic as shown in Figure 7. Now we take this desired sideband located at *B* in Figure 6 and put it into a second modulator where we modulate a second frequency of about 89,200 cycles. There will then be produced two new sidebands, one shown at *D* running from 56,000 to 58,700 and one shown at *E* running from 119,700 to 122,400. The new *D* and *E* sidebands are very far apart and also 30,000 cycles removed from the second carrier, and it becomes a very easy matter to build a filter which selects the desired band *D* and discriminates against the 89,200 cycle carrier and sideband *E*. This filter does not have to have anywhere near the steepness of attenuation slope that the first one does because of the relatively greater separation between the bands and the carrier *K*.

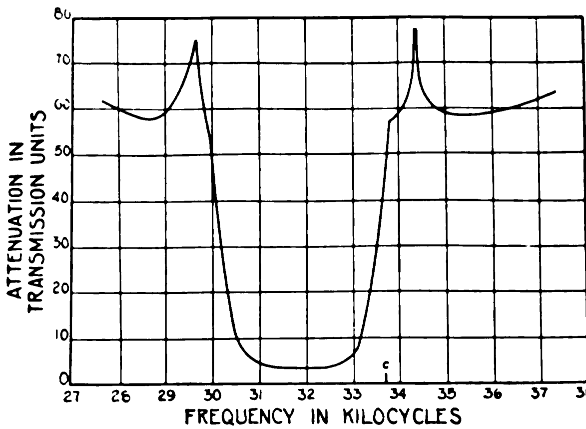


FIGURE 7—Attenuation Curve of the First Filter

By this double modulation process we also provide ourselves with a flexibility in frequency range which we could not attain by the simple scheme except at prohibitive expense. That is, if we build our second filter to transmit frequencies between 71,000 and 41,000 cycles we can cause our desired band *D* to fall anywhere within this range such as *D′* by merely moving the second carrier *K* to *K′*. If that carrier is removed down to 74,000, the lower

sideband then falls between 41,000 and 44,000. If we move the carrier K up to 101,000, the sideband runs from 68,000 to 71,000. We thus secure a flexibility in frequency range for the placing of our sideband D with the use of fixed frequency band filters, which, for work such as we have been doing, is of vital importance.

The question may be asked why we picked the lower sideband at 33,700 and used it to modulate another frequency and then again picked the lower sideband. The reasons for this are partly circuital and partly psychological. We could have picked the upper sideband at 33,700 and then modulated about 93,000 cycles and located a sideband in the same region where we have D as represented. In that case the sideband would be reversed. There is no electrical reason for desiring the band as we have used it, over reversing the band, as either will give just as good quality, but it seemed simpler to maintain the frequency arrangement in the same order in which it occurs in the voice. There is an objection to producing the sideband D by using one of the sidebands near 33,700 to modulate a second carrier of about 21,800 which would again place the sideband D in about the same position. The objection here lies in the fact that there is some likelihood of harmonics, especially second harmonics, giving some trouble if the balance is not perfect. It seemed desirable in a first experimental installation to keep all the frequencies and bands totally separate and not have them overlapping in such a way as possibly to give rise to any harmonic trouble. We, therefore, chose the lower sideband in both cases, which altho it means turning the frequency band over twice, yet finally places it in the desired position and gives us the flexibility which is of value.

Replacing the Carrier

At the receiving station it is necessary to replace the carrier It is not necessary to replace the auxiliary carriers used at the transmitting station: 33,700 and 89,200, but only the resulting or final carrier 55,500. It is interesting to note that this final carrier which is "eliminated" is not generated at the transmitting station at all. It is generated only if the first modulator is unbalanced and some of the first carrier gets into the second modulator. In practice the carrier is considered eliminated if reduced in amplitude to a few percent of its original value.

The accurate replacing of the carrier is sometimes of great importance. This is particularly true in receiving music, as other-

wise overtones would not be overtones at all. As far as receiving speech goes, if the carrier is placed too close to the sideband, the voice sounds low and gutteral, while if placed too far away, it appears very high pitched, but in either case the articulation is reduced from what is secured when the carrier is correctly placed. It is, therefore, necessary for satisfactory operation to place the carrier as near as possible to the theoretical point.

If our carrier is to remain within say 20 cycles of the theoretical point, that means that both the suppressed carrier and the replaced carrier must remain constant within 10 cycles. If our carrier has a value of say 55,500 cycles and we wish to keep the frequency within 10 cycles, that means that it has to stay within 1/55 of a percent of the desired value at all times even tho temperatures in the room change or the voltage supply fluctuates slightly. To secure this constancy is a job all by itself. Ordinarily an oscillator changes its frequency when either the plate voltage or filament voltage changes, or when the temperature changes affect the constants in the circuit, and steps had to be taken to prevent these changes or minimize the effects.

Present System

The system which we have in use at Rocky Point is outlined schematically in Figure 8 and the circuit is given in Figure 9.

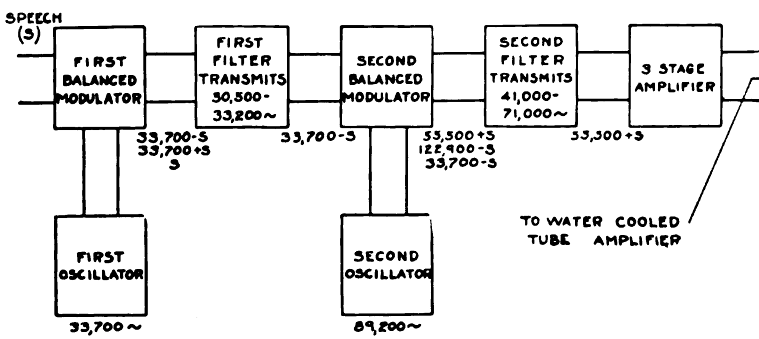


FIGURE 8—Schematic Arrangement of the Constituent Elements in the Single Sideband Apparauts

The first carrier is 33,700 cycles. As previously stated, a relatively low frequency was chosen to give us good elimination of the undesired sideband. The oscillator generating this frequency is usually referred to as the first oscillator. The first carrier is impressed upon the grids of the balanced modulator tube by means of two transformers. It is impressed on the two grids in the same

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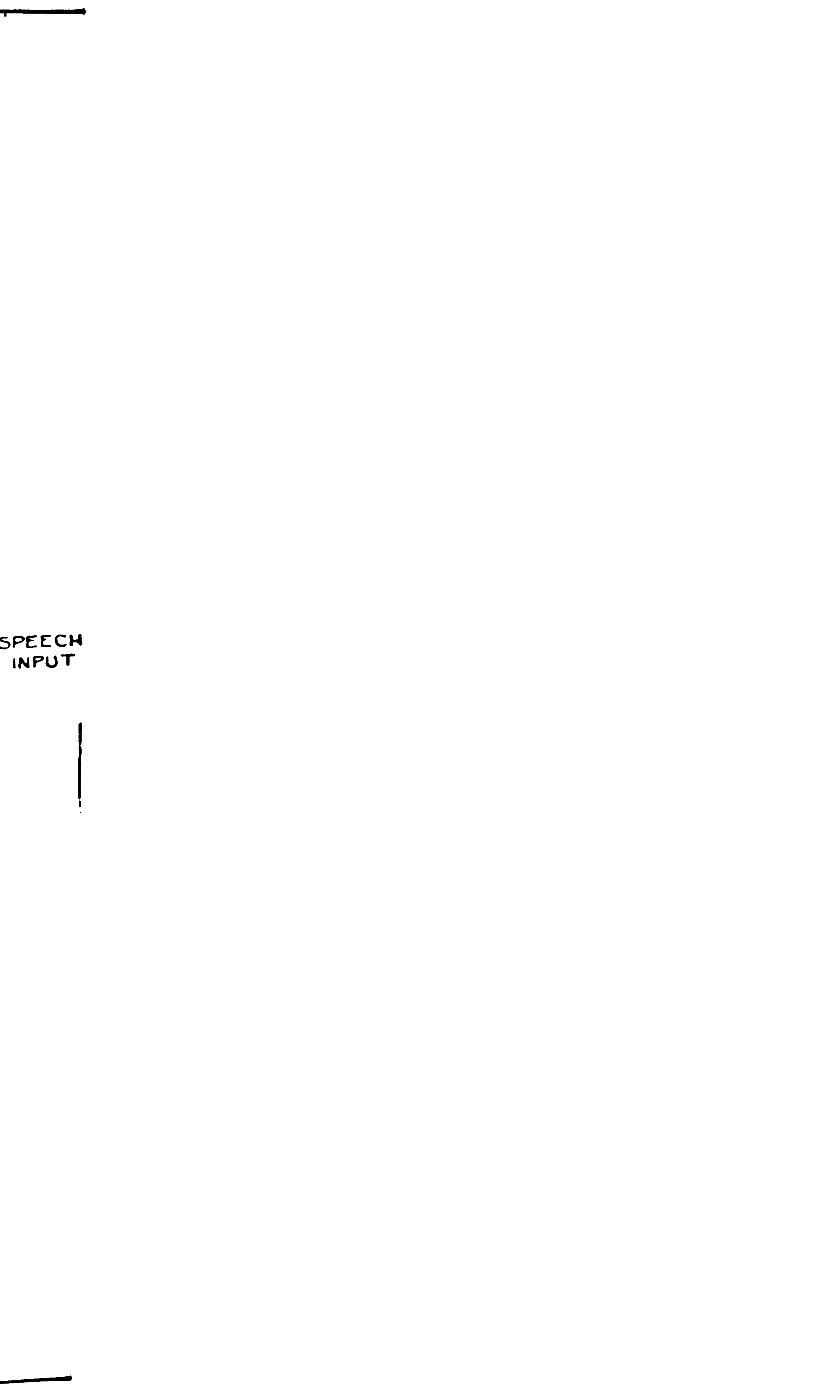

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phase (equivalent of position A, Figure 5) so that when we use the differential transformer in the plate circuit, the carrier is largely eliminated. The speech comes in from the line or from an amplifier, and is impressed on the grids of the modulating tubes in opposition (position B in Figure 5). In the plate circuit of the modulator the differential transformer passes the sidebands and the speech. By using a transformer here which is inefficient for the speech frequencies but efficient for the sideband, the speech frequencies will be discriminated against. The sidebands with a small amount of the signal frequencies then go into the first filter. The transmission characteristic of the first filter is shown in Figure 7. Its impedance characteristics are shown in Figure 10. This filter, tho having the theoretical cut-off at 33,200 does not cut-off sharply. The attenuation begins to increase rapidly at that point, so that frequencies 500 or 600 cycles higher are not entirely eliminated but are reduced somewhat in amplitude.

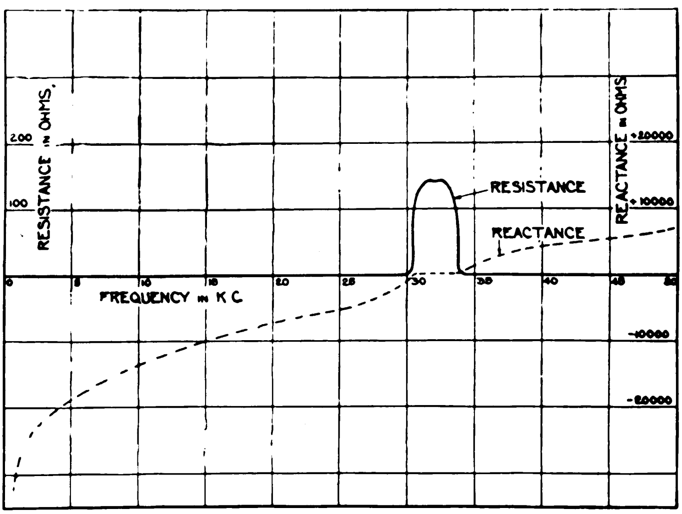


FIGURE 10—Impedance Curves of the First Filter

From the filter the single sideband passes to the second modulator. This second modulator is also of the balanced type in order to reduce the amplitude of the second carrier and not overload the filter. The second carrier is supplied from the second oscillator in Figures 8 and 9 which operate at about 89,200 cycles. The modulating frequencies now run between 30,500 and 33,200 and are impressed in opposite phase on the grids of the two modulating tubes. The 89,200 cycle carrier is impressed

on the two grids in phase. The transformer in the output is differentially connected so as to eliminate the carrier, but it transmits the sidebands and the modulating frequencies.

The two sidebands pass from the modulator into the second filter. The transmission band of the second filter is from 41,000 to 71,000. The attenuation curve for it is shown in Figure 11. Only one of the sidebands falls within this range. The impedance characteristic of this filter is shown in Figure 12. It will be observed that the impedance of this filter is quite low in the neighborhood of 30,000 to 33,000 cycles. The resistance outside the band, of course, is practically zero and the reactance curve crosses the axis at this point. The purpose of using a filter having this reactance characteristic is to allow the modulators to function properly, as it is a well-known fact that in order to get modulated power out of the Van der Bijl type of modulator, the impedance in the plate circuit for both the modulating and modulated frequencies must be low. The impedance is, therefore, made a minimum for the modulating frequencies which, in this case, lie between 30,500 and 33,200 cycles. It is not necessary to make it either zero or a minimum for the modulated frequency of 89,200 in the arrangement which we are using, for the reason that the differential connection of the transformer eliminates the filter from the circuit. In the case of the first modulator, the same requirements hold. The differential transformer connection eliminates the filter from the circuit for the carrier frequency, but not for the speech frequencies. The filter used has not the desired low impedance at the speech frequencies, so we take advantage of the inefficiency of the transformer at these frequencies to provide the low attached impedance.

The transmission characteristics of the second filter are such as to give us considerable flexibility in frequency range. The lowest position where it will allow the placing of the desired sideband is between 41,000 and 44,000 cycles. The frequency of the second oscillator in this case would be set at 41,000+33,200 or 74,000 cycles. This point is well up on the upper side of the attenuation curve, so that the second carrier frequency would be kept out of the amplifier and the antenna. The degree to which this muct be kept out is very great for the reason that in a high-powered set, it does not take a very large input to put several watts into an antenna even tho it is off tune. The highest position where we would possibly place the desired sideband is around 68,000 to 71,000 cycles. The frequency from the second oscillator would then be about 101,000 cycles. The upper sideband in

both of these cases would be at 100,000 cycles or above. The second filter will easily eliminate this upper sideband.

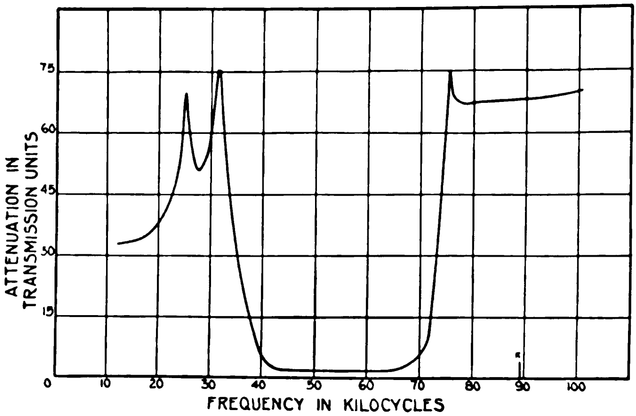


FIGURE 11—Attenuation Curve of Second Filter

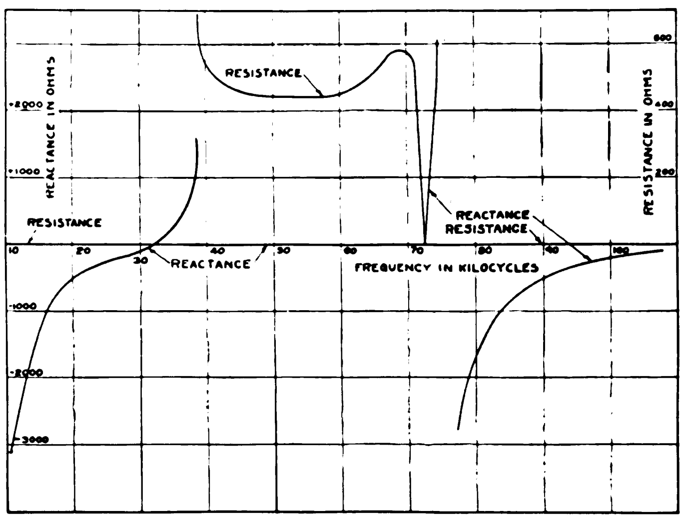


FIGURE 12—Impedance Curves of Second Filter

The second filter is built to have a very high attenuation between 24,000 to 35,000 cycles, because it is in this region that the modulating frequencies of the second modulator lie. The arrangement of the second modulator is such that the modulating frequencies pass thru both tubes and the second transformer readily into the second filter so that it is not impossible, if the second filter does not eliminate them, for them to produce in suc-

ceeding amplifier tubes second harmonics which might lie directly in the range of the desired sideband. That is, since our first sideband lies between 30,500 and 33,200, the second harmonics from it would lie between 61,000 and 66,400, and if we try to use this latter region as the position of a sideband for communication, we might find that some of these harmonics would fall within the band and give disturbing noises.

From this second modulator the desired sideband *D*, which is shown in Figure 6, is passed into the low power amplifier. This low power amplifier is a three-stage amplifier consisting of 5-watt, 50-watt, and 750-watt steps. These are the power ratings of the tubes. The actual power secured in the various amplifier stages is not these values, but is considerably lower. Power efficiency in this part of our set is not of importance, but quality is, so these three stages are built for reproducing the desired sideband faithfully and at a sacrifice of power. The last two stages are purely voltage step-up stages, or choke coil amplifiers. The power secured from the last amplifier is about 500 watts maximum.

Installation

Photographs of the single sideband apparatus are shown in Figure 13. The apparatus is built on two racks. Each elemental circuit is also on its own panel. The first rack contains all the single sideband producing apparatus and the second rack contains the three-stage amplifier with the testing and measuring panels.

The power comes from several sources. The modulators and oscillators have their plates supplied from the 220-volt direct current circuit in the station. The amplifiers are supplied from a 1,500-volt generator. All filaments are lighted by alternating current. The negative grid potentials for the amplifiers are secured by potentiometer arrangements from the 220-volt circuit while for the modulators, a battery is used.

The arrangement of this apparatus at Rocky Point is shown in Figures 14 and 15. Figure 14 shows the single sideband producing rack and Figure 15 shows the preliminary amplifier rack. In locating this apparatus in the station, precautions had to be taken to prevent singing. The power supplied from either oscillator to the modulators is of the order of one one-thousandth of a watt. The power delivered by the water-cooled tubes to the antenna runs up over 100 kilowatts. The ratio of these powers is about 1 to 100,000,000. It would not do to leave this apparatus operated by such small voltages in such a position that the high-

powered equipment could disturb it. This apparatus was, therefore, all mounted inside a copper-screened cage. The screen was placed on the floor and ceiling as well as on all four sides. Even

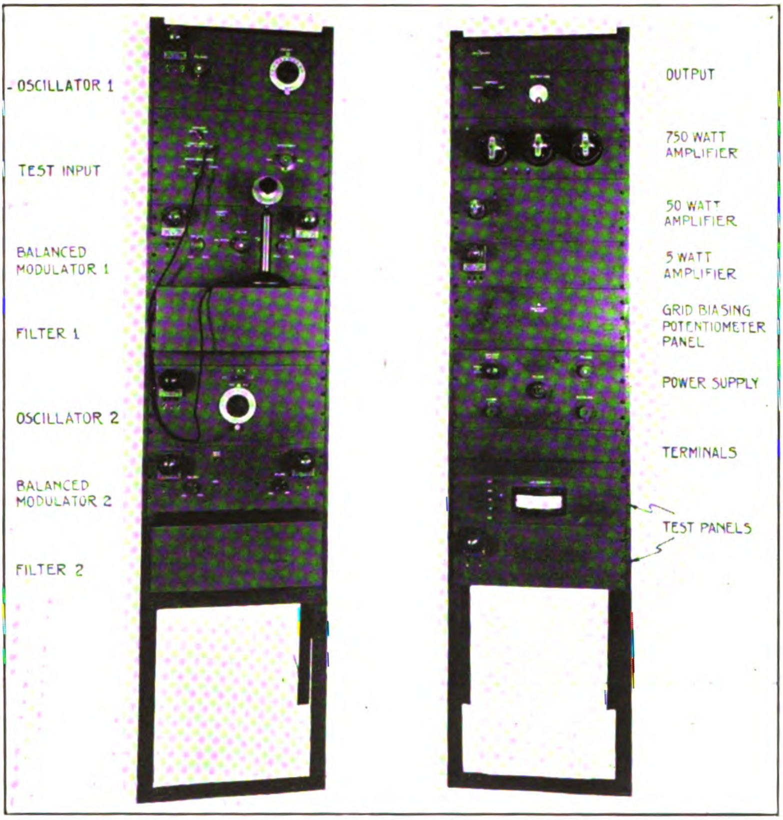


FIGURE 13—Photograph Showing the Panel Type of Construction. The Left Hand Rack Contains the Oscillators, Modulators, and Filters. The Right Hand Rack Contains the Three-stage Amplifier which Delivers the Single Sideband at About 500 Watts Maximum

a screen door was provided, tho it has not always been necessary to close the latter. Shielding is sufficiently well done, so that any voltages introduced in the wiring of the set from the high-powered apparatus are small compared to the driving voltages from the oscillators.

There is also located in this cage a Vreeland oscillator which provides frequencies over the audio range for much of the test work. It may be observed in Figure 14 behind the sideband rack.

The power supply of all this equipment was handled thru a

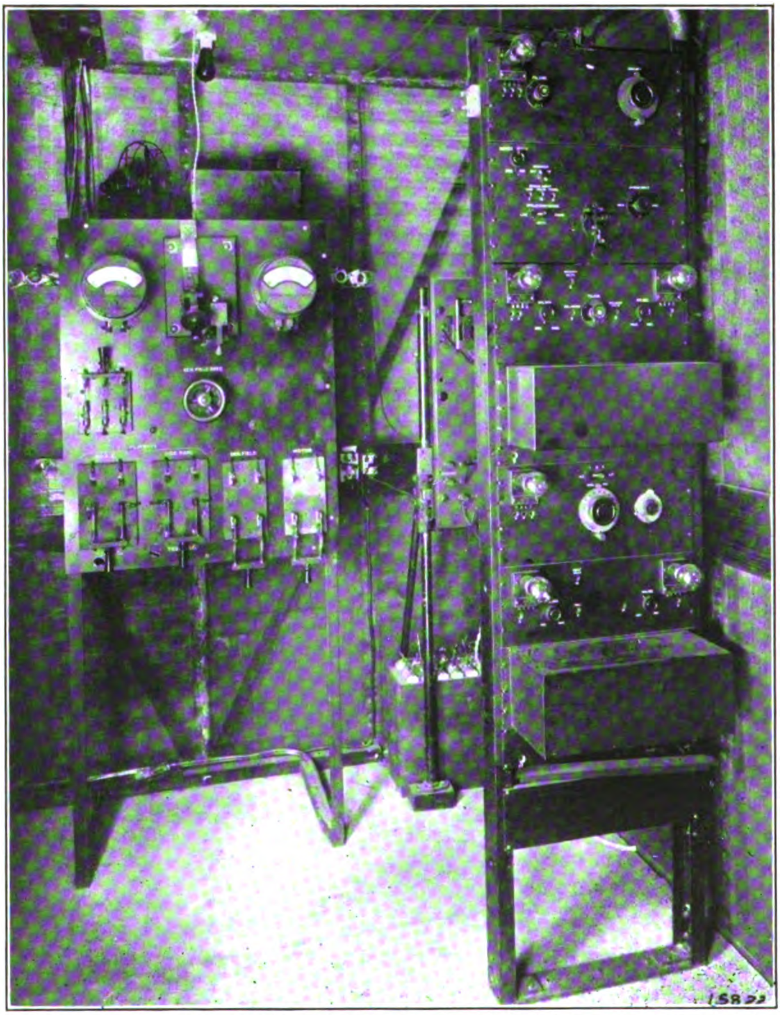


FIGURE 14—Photograph Showing the Oscillator-Modulator Rack Located in the Shielded Cage

switchboard also located within the cage. The motor-generator which supplies the power at 1,500 volts for the three 250-watt tubes is started and controlled from this panel. The direct current power circuits for supplying the 220 volts to the small tubes is also run thru this switchboard. Other pieces of apparatus used in testing such a wave meter and monitoring receiving set, are also usually kept in the cage tho they are not shown in any of the pictures.

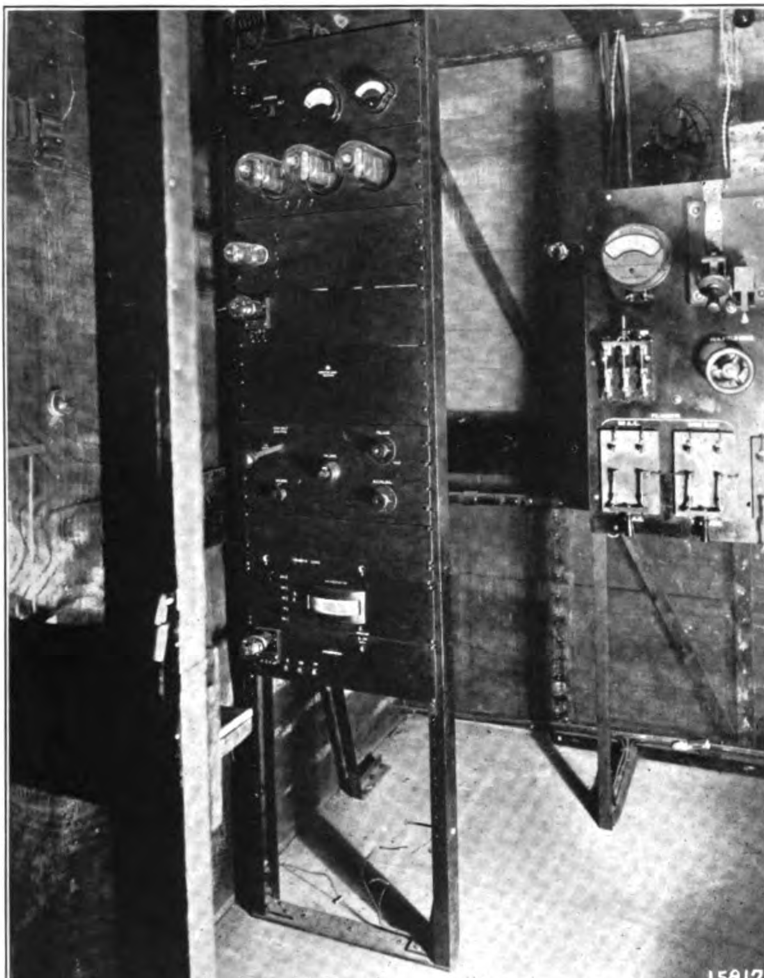


FIGURE 15—Photograph Showing the Preliminary Amplifier Rack Located in the Shielded Cage

PERFORMANCE

In a study of the performance of the single sideband apparatus the first element we look for is quality. We get our idea of quality primarily from an amplitude-frequency performance curve. This is based upon our previously stated theory that if all sustained frequencies between 200 and 2,500 cycles are transmitted without any appreciable frequency discrimination, the quality will be satisfactory for the purpose. Our quality tests, therefore, take the form of a set of curves plotted between input frequency in the

audio range and output amplitude from the last amplifier. The amplitudes of the input frequencies are kept constant, that is, we supply the same power at all audio frequencies. The voltage or current of the single sideband resulting is measured for each one of the signal frequencies and the curve plotted. We also find it desirable to take measurements at various points in the set in order to locate the position of the various distortions if possible. In Figures 16 to 18 are a set of these curves which were taken in the manner described.

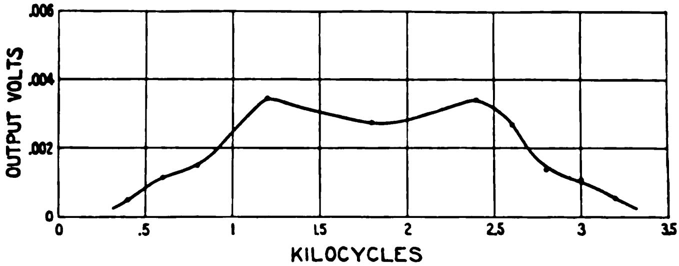


FIGURE 16—Amplitude of the Sideband Frequencies at the Output Terminals of the First Filter as a Function of Modulating Frequency. Input at All Frequencies Constant

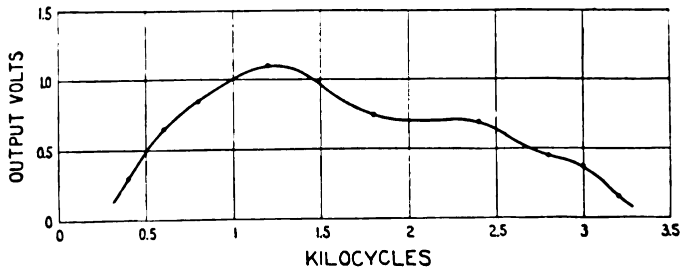


FIGURE 17—Amplitude of the Sideband Frequencies at the Output Terminals of the Second Filter as a Function of Modulating Frequency

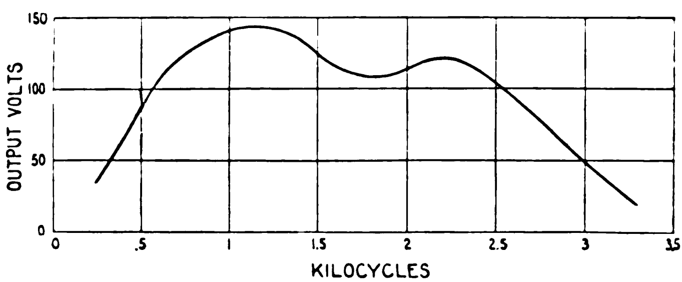


FIGURE 18—Amplitude of the Sideband Frequencies Delivered by the Set to the Water-cooled Amplifying Tubes as a Function of Modulating Frequency

Figure 16 is a curve showing the output of the first modulator and filter. In Figure 17 is shown the output curve of the second filter. Some additional distortion has evidently occurred over what is produced in the first one. However, it does not produce any serious reduction in quality. In Figure 18 the over-all characteristic is shown. Further distortions occur, some parts being worse and some parts better. This curve is still one which indicates we should get adequate quality.

That the quality resulting from the set is good is indicated by the fact that in the public demonstration across the Atlantic, the speakers' voices were recognized and reporters had no trouble in getting every word using the head telephones or a loud speaker. We have received word from some nearby listeners who said the quality was not good, but their troubles were located in their sharp receiving circuits. The distortion that a good long wave telegraph receiver will cause is enormous. When this fact was pointed out and proper circuits used, their bad quality disappeared.

The operation of this apparatus has been quite satisfactory. It has been in use for a year and a-half. It was operated continuously during the early trials and development and during the last year during all weekly tests. Changes have been made from time to time, as can be seen if Figures 13 and 14 are compared. Photograph 13 was taken just before the apparatus was shipped to Rocky Point and Figures 14 and 15 after it has been in use some time. The continued operation could not help but cause certain modifications to be made to improve operation or facilitate adjustment or control. All changes made, however, were of a minor nature, as no departure was made from the fundamental system which we had in mind when starting out. The operation is reliable in every way, as evidenced by the regular week-end trans-Atlantic tests which are carried out and the absence of necessity of tinkering between times.

Research Laboratories of the American Telephone and Telegraph Company and Western Electric Company, Incorporated, New York.

October 9, 1924.

SUMMARY: This paper describes in detail the equipment and circuit used in the production of the single sideband for trans-Atlantic radio telephony in the experiments at Rocky Point. The set consists of two oscillators, two sets of modulators, two filters, and a three-stage amplifier. The oscillators and modulators operate at power levels similar to those in high-frequency communication on land wires. The three-stage amplifier amplifies the sideband

produced by these modulators to about a 500-watt level for delivery to the water-cooled tube amplifiers.

The first oscillator operates at about 33,700 cycles. The modulator is balanced to eliminate the carrier; and the first filter selects the lower sideband. In these trans-Atlantic experiments the second oscillator operated at 89,200 cycles, but might operate anywhere between 74,000 and 102,000 cycles. The second modulator, which is also balanced, is supplied with a carrier by the second oscillator and with modulating currents by the first modulator and first filter. The second filter is built to transmit between 41,000 and 71,000 cycles, so that by varying the second oscillator, the resulting sideband, which is the lower sideband produced in the second modulating process, may be placed anywhere between these two figures. Transmission curves for the filters are given as well as some amplitude-frequency performance curves of the set.

POWER AMPLIFIERS IN TRANS-ATLANTIC RADIO TELEPHONY*

By

A. A. Oswald

and

J. C. Schelleng

(Western Electric Company, New York)

Introduction

It has long been recognized by engineers interested in telephone service over great distances that radio transmission offers the most promising medium thru which commercial trans-oceanic voice communication may be realized. It is now fairly well established that, both from the viewpoint of the conservative use of the ether and from the consideration of financial costs, the most economical radio system requires the development of the useful signal element by modulation and filtration at very low power values and subsequent amplification to the levels necessary for successful transmission. Convincing proof of this statement is found in the successful application and operation of such a system during the trans-atlantic telephone experiments in 1923.

It is not within the scope of this paper to enter into a complete discussion of these experiments or even to describe the operation of the system as a whole. Papers have already been published[1] which present the over-all results attained during recent tests and which describe the general system and methods employed.

For the present purpose it will suffice to say that one of the objects of the experiments in 1923 was to demonstrate the efficacy and the practicability of radio transmission by means of the single-sideband eliminated-carrier method. With this method

*Presented before The Institute of Radio Engineers, New York, May 7, 1924. Received by the Editor, December 11, 1924.

[1] "Trans-oceanic Wireless Telephony," by Dr. H. W. Nichols, "Journal of the Institution of Electrical Engineers," volume 61, number 320, July, 1923 "Transatlantic Radio Telephony," by H. D. Arnold and L. Espenschied "Journal of American Institute of Electrical Engineers'," August, 1923.

the narrowest possible frequency band is employed in the ether and all of the radiated energy has maximum effectiveness in transmitting the signal. As mentioned above, the single sideband currents of the desired frequency are prepared at low power and then amplified to the required magnitude for application to the transmitting antenna. A paper has been given recently which described the low-power modulating system in considerable detail.[2] The present paper deals entirely with the power amplifiers. Material concerning other parts of the system will only be introduced when it contributes to the definition of the amplifier functions or when it seems essential to a clear conception of the requirements imposed on the amplifier.

Characteristics of Single-Sideband Signal

It is well known that, when an alternating carrier current is modulated by telephone currents, the resultant wave is distributed over a frequency range[3] which may be conveniently considered in three parts: (1) the carrier frequency itself, (2) a frequency band extending from the carrier upward, and having a width equal to that of the frequencies appearing in the modulating wave, and (3) a band extending from the carrier downward, and having a similar width.

These relationships are shown by the three spectra plotted in Figure 1. The rectangle B with the frequency limits S_1 and S_2 represents the voice frequency band essential to the transmission of intelligible speech. The frequency f_c is that of the alternating carrier current which is being modulated. The products of modulation are spread over a region comprising the original carrier frequency f_c and two bands B_1 and B_2, known as the lower and upper sidebands. These two bands have the same width as the band B and furthermore each transmits power which contains all of the elements necessary to reproduce the original speech. Since most of the radio telephone transmitters now in use radiate all of the products of modulation shown in Figure 1, it will be convenient to employ the term "ordinary" to designate systems in which none of these products are suppressed.

The ordinary radio telephone sends out the carrier f_c continuously and adds to this the bands B_1 and B_2 whenever speech or other similar signals are transmitted. The total transmission

[2] By R. A. Heising, entitled, "Production of Single Sideband for Trans-atlantic Transmission," presented at a meeting of The Institute of Radio Engineers, on March 19, 1924.

[3] "Relations of Carrier and Sidebands in Radio Transmission," by R. V. L. Hartley, Proceedings of The Institute of Radio Engineers, February, 1923.

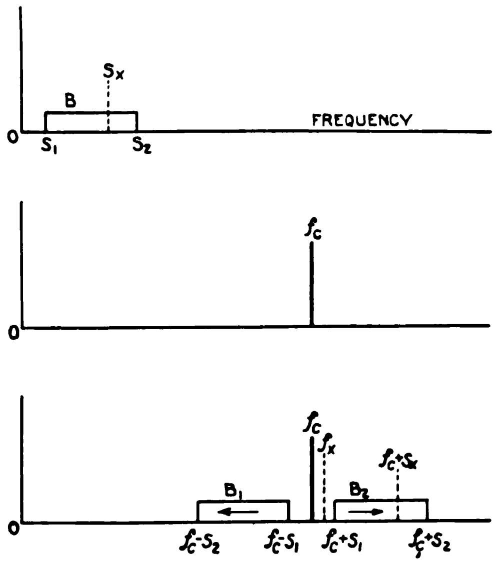


FIGURE 1—Frequency Spectra in Radio Telephone Modulation

frequency range extends from $f_c - S_2$ to $f_c + S_2$ and is therefore equal to 2 S_2.

In the single-sideband eliminated-carrier method of transmission no power is radiated except when signals are being transmitted. A combination of modulators and electrical filters are employed to eliminate the carrier and one sideband. Thus in Figure 1, currents of all the frequencies less than f_x are prevented from reaching the output circuit of the modulation system. The total transmission frequency range extends from $f_c + S_1$ to $f_c + S_2$ and is therefore equal to $S_2 - S_1$. This is slightly less than one-half the frequencies employed by the ordinary method. Modulation becomes a simple frequency transformation in which all the signal frequencies in the band B are stepped up an equal amount f_c to produce the band B_2.

If a pure sine wave potential of frequency S_x be applied to the input terminals of a single-sideband system, the resultant output will be a continuous wave of frequency $f_c + S_x$. Hence a single-sideband radio telephone system can be converted into a continuous wave telegraph system without any change in apparatus other than replacing the microphone with a single frequency generator.

Thus far the comparison of the single sideband signal with that of ordinary radio telephone signal has emphasized the

difference in frequency ranges. There are other important differences between the two signals which have a direct bearing on the requirements of the transmitting apparatus. Such differences will now be considered with the assistance of the diagrams shown in Figure 2. In these diagrams no effort has been made

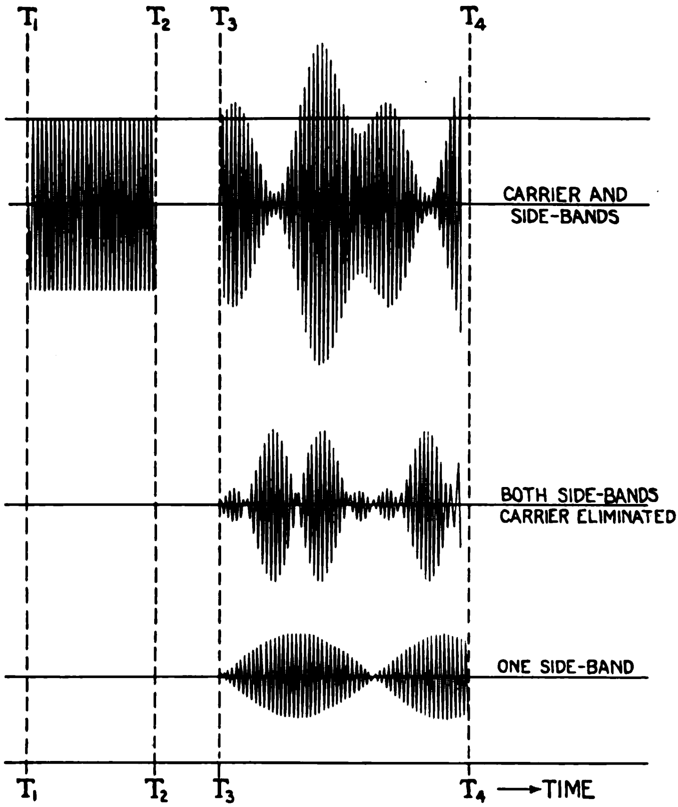


FIGURE 2—Wave Radiated in Three Systems of Transmission. In interval T_1—T_2 there is no modulation. In T_3—T_4 there is modulation by two audio frequencies, a fundamental and its second harmonic

to represent correctly each radio frequency cycle, but the envelopes of the radio frequency instantaneous peak values are properly related. The signal condition shown during the time interval from T_1 to T_2 is that which occurs when no voice message is being transmitted, for example, during a pause between words. The signals shown during the interval from T_3 to T_4 represent conditions when a voice message is in process of transmission and the audio frequency wave conforms to the envelope of the

upper curve. The signals represented by the upper curve are those radiated by the ordinary radio telephone transmitter. In this case a continuous wave signal is sent out during the interval $T_1\ T_2$ and the wave, transmitted during the interval $T_3\ T_4$, may vary in amplitude from zero to twice the amplitude of the continuous wave signal radiated during the interval $T_1\ T_2$. The signals represented by the lower curve are those sent out by a single-sideband radio telephone transmitter. No signal of any sort is radiated during the interval $T_1\ T_2$ and the maximum amplitude during the period $T_3\ T_4$ is considerably less than that for the case where the carrier is transmitted. Even disregarding the advantages possessed by the single-sideband system, that there is less interference due to the sharper tuning permissible in the receiver and that the received signal strength is subject to less variation due to changes in the ether conditions, this system has the advantage of lower power consumption for given results. The maximum power capacity requirement is one-fourth of that for the usual transmitter. It is important to remember that the power output of a single-sideband transmitter is zero when no speech is transmitted and that the output varies from zero to approximately full load each time a word is spoken. This characteristic has an important bearing on the power supply and amplifier requirements.

Operational Classification of Amplifiers

Vacuum tube amplifiers, like most electrical apparatus, may be classified in a number of ways. For the present purpose it will be convenient to group them in three classes, according to their mode of operation, defined as follows:

Class I comprises amplifiers in which operation is confined to the substantially linear portion of the tube characteristic curve.

Class II consists of those in which the anode current never ceases to flow, but operation extends beyond the linear portion of the tube characteristic.

Class III comprises amplifiers in which the anode current ceases to flow during a portion of each cycle.

Telephone repeaters and similar audio frequency apparatus, in which distortionless amplification is a prime consideration, are usually amplifiers of the first class. When some distortion is permissible and power efficiency is important, amplifiers of the second and third classes are employed. High efficiency radio frequency power amplifiers are all of the third class.

Typical anode current and potential relations for each class

of amplifier are shown in Figure 3. The curves at the left show the static characteristic and also the dynamic characteristics for a load at unity power factor. Those at the right are the corresponding current and voltage relations with time as abscissa when a simple sinusoidal voltage is applied to the grid. In an

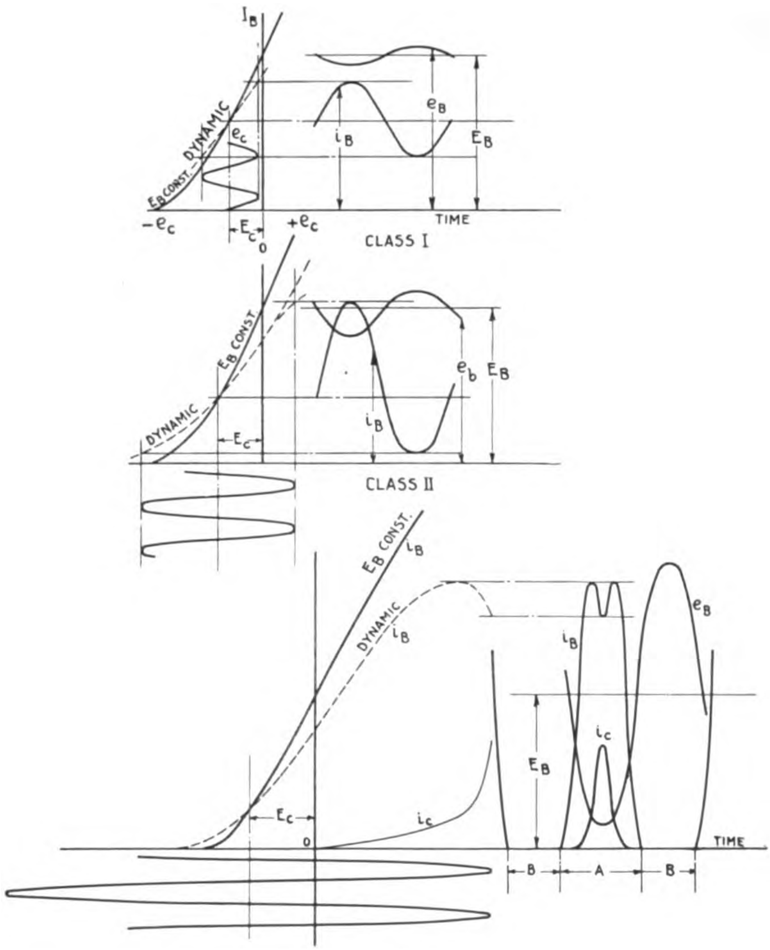


FIGURE 3—Typical Characteristics of the Three Classes of Amplifiers

amplifier of the first class the grid polarizing potential E_c is selected so that the alternating emf. applied to the grid operates on the linear portion of the dynamic characteristic. The peak value of the applied alternating grid voltage e_c is always less than the polarizing potential E_c. The alternating components of

anode potential and current very closely approximate sine waves and may be considered so for all practical purposes. In the case of amplifiers of the second class, the peak value of applied alternating grid potential may exceed the grid polarizing potential, but is never sufficient to bring the anode current to zero. The anode current i_b departs considerably from a sine wave. Distortion may occur in i_b due to the flow of grid current when the grid assumes positive values. Amplifiers of the third class are driven by comparatively large alternating grid potentials. The value of grid polarizing potential is selected at or near the point of anode current cut-off. For some purposes it may be made much greater than the cut-off value. It is usually the case that during each cycle the grid swings positive to such an extent that there is an appreciable grid current. The anode current is pulsating and flows only during the time A. The energy dissipated in the anode is represented by the integrated product of e_b i_b. Since i_b is zero when e_b is large, this mode of operation results in much higher power efficiencies than can be obtained otherwise. While discussing this question of efficiency it may be pointed out that for amplifiers of the third class a considerable advantage is gained by limiting the flow of anode current to a half cycle or less. This condition is obtained by establishing the grid polarizing potential at values equal to or greater than the value required for anode current cut-off and exciting the grid with correspondingly large values of alternating emf. Improved efficiency may be obtained in this manner for the case of continuous wave telegraphy and that of carrier radio telephony without serious disadvantage in other respects. However, in the case of single-sideband transmission other considerations make it desirable to employ grid polarizing potentials slightly less than the anode current cut-off value. The foregoing definition of third class amplifiers includes those in which the grid and anode potential wave form are distorted for the purpose of obtaining high efficiency or high power. The essential characteristic common to all members of this class is that the current becomes zero for a finite portion of the cycle.

Amplifiers of the Third Class Applied to Radio Telephony

Since the anode current in an amplifier of the third class is pulsating, energy is delivered to the output circuit in similar form. Therefore, it is usually advantageous to employ an output circuit containing elements which tend to store the energy as received from the tube and power source and to deliver this

energy to the load in sinusoidal form. Furthermore, altho the presence of harmonics in the anode circuit contributes to the attainment of high efficiencies, it is obvious that, where a third class amplifier is used as the final stage of a radio telephone system, the radio frequency harmonics must not be allowed to reach the antenna. The importance of harmonic current suppression in a high power radiating system can scarcely be over-emphasized. One of the best ways to accomplish both the foregoing results is to employ a circuit which by-passes harmonics thru a low impedance path and, by virtue of its resonance to the fundamental, receives energy at that frequency. Such a circuit is shown in Figure 4.

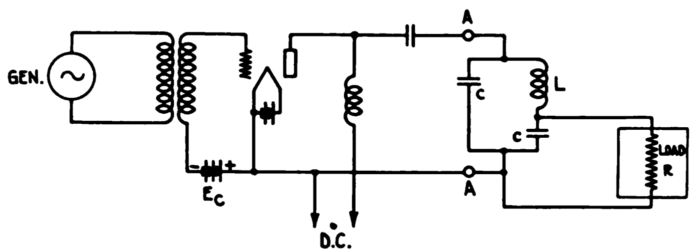


FIGURE 4

During a portion of each cycle the circuit CL, which is tuned to the frequency of the driving generator, receives one or more current pulses. These impulses add energy to that being cyclicly interchanged between the electrostatic fields of the capacities and the magnetic fields of the inductances. The movement of energy between C and L continues during the period when the anode current is zero, and the alternating potential across the terminals AA of CL completes each cycle approximately as a sine wave. It follows that the load current is not greatly affected by the pronounced distortion of the anode current.

Referring to Figure 5, e_c is the alternating emf. applied to the grid, and i_b is the anode current. In this case the grid polarizing potential E_c was selected to cut-off the anode current at $e_c = 0$. The anode current consists of distorted half waves, but the circuit CL, Figure 4, functions to give the load current a sinusoidal form. The anode current, Figure 5, may be resolved into a direct current component, a radio frequency fundamental and a large number of harmonics.

Since the dynamic characteristic is a function of input amplitude[4] the behavior of the system is somewhat different when a

[4] R. A. Heising "Modulation in Radio Telephony," (Appendix), PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, August, 1921.

modulated radio wave is applied to the grid, but the underlying principle of operation is unchanged. The variations in dynamic characteristic change to some extent the radio frequency distortion in the anode circuit.

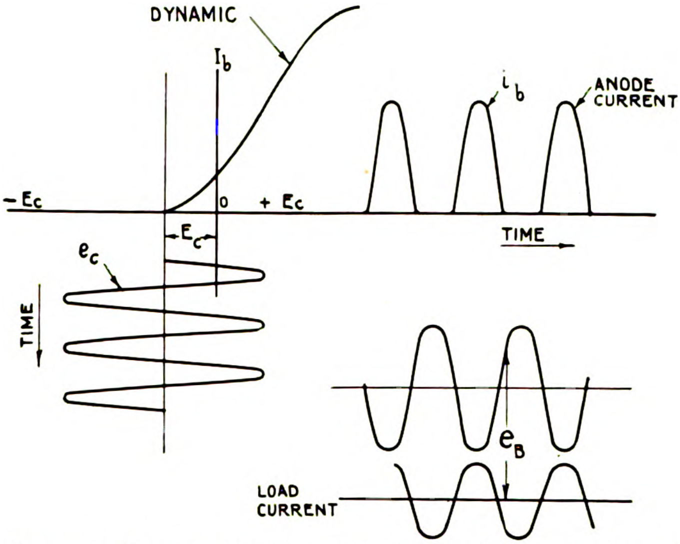
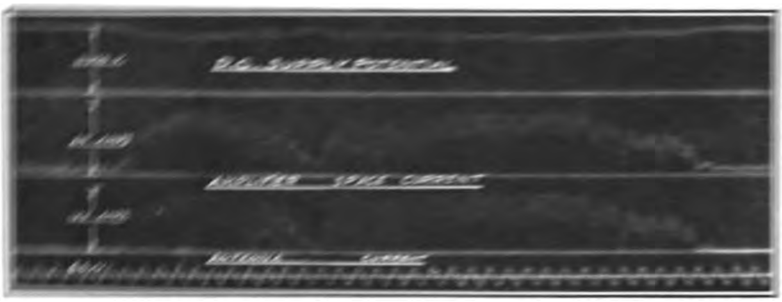


FIGURE 5—Current and Voltage Relations in Typical Amplifier of the Third Class

When a modulated radio wave such as the upper curve in Figure 2 containing the carrier and both sidebands is amplified, the anode current contains a constant direct current component corresponding to the carrier and a large number of audio frequency components corresponding to the original voice currents. Provision is usually made to by-pass these voice currents around the direct current power source. The load on the power source fluctuates at syllable and word frequencies within a relatively narrow margin at approximately full load value.

On the other hand, when a radio wave comprising a single-sideband signal, such as the lower curve in Figure 2, is amplified, the anode current does not contain a constant direct current component in addition to the radio frequency currents, but does have large currents within the voice frequency range, plus large low frequency components corresponding to words and syllables. Hence the load on the power source fluctuates between zero and full load at syllable frequencies. This is an important consideration in designing the power source for a single-sideband amplifier.



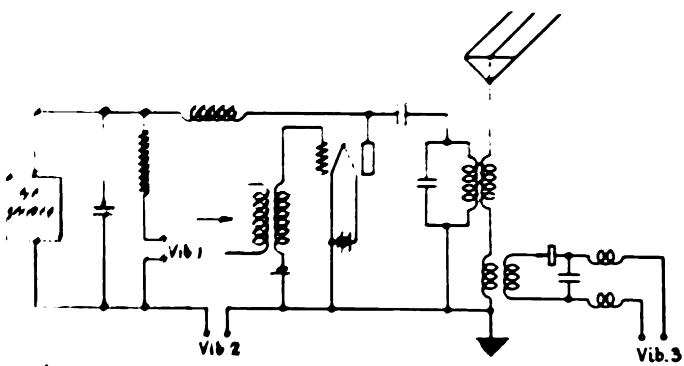


Figure 7 — Connection of the Oscillograph Vibrators Used in Taking Figure 6

special precautions are taken to eliminate them from the antenna. One well-known method of balancing out the even harmonics is by means of the push-pull amplifier. The odd harmonics, however, are not affected and hence the method of itself is not sufficient, altho it is helpful. In a radio transmitting system this reduction of the even harmonics is the main virtue of the push-pull amplifier. As regards the reduction of distortion in the speech signal which is being transmitted, little if any advantage is gained. When the output circuit is such as to have negligible impedance for the predominant harmonics, there is no difference whatever between the amplitude characteristics of these two types of amplifier.

In the case of the present trans-atlantic experimental installation, the introduction of a push-pull amplifier presented certain minor problems. While these were by no means difficult to solve, the advantage to be gained did not justify discarding the more simple method of connecting all of the power tubes in parallel.

Water-Cooled Tubes

At the time when the decision to build a trans-atlantic telephone transmitting set was made, a suitable vacuum tube had already been developed in this laboratory. The principles which have led to the adoption of the type employed have already been fully discussed by Dr. W. Wilson in the "Bell System Technical Journal," volume 1, number 1.

For those who have not followed recent progress in the development of tubes of high power it will be sufficient here to say that in this type the anode, which forms a part of the containing envelope, is immersed in water. Continuous circulation of the latter carries away the power dissipated in the copper anode, the temperature of which is thus maintained well below the boiling point of water. Instead of the tube being inserted in a socket it is placed in a jacket so designed that water, entering from below, flows about the cylindrical anode with a rapidly whirling motion.

The type of tube employed in the experimental installation is capable of dissipating continuously ten kilowatts with an ample factor of safety. Such a tube will develop ten kilowatts in a suitable oscillating circuit. The operating voltage is 10,000. The ten-kilowatt output can readily be obtained with a direct current in the anode circuit of 1.4 amperes or less. The grid current is between 0.1 and 0.3 amperes. As a rule it is desirable that the filament emission of an efficient radio power tube be

about five times the direct current to the anode. This means that in the present case about seven amperes are needed, a requirement which is readily met with a filament current of 41 amperes at 22.5 volts. Pure tungsten wire 0.035 inches in diameter is used. The grid must usually be able to dissipate from 200 to 250 watts. This condition is fulfilled by the water-cooled tube with a large factor of safety. The amplification constant is about thirty-eight. This signifies that the negative potential which we must apply to the grid in order to reduce the anode current at 10,000 volts to zero, has a value of about 260 volts, that is $10{,}000/\mu$.

When large numbers of high power tubes are operated in parallel it is desirable that separate protection be provided for each one. If this is not done, an open grid circuit, a burned-out filament or some other abnormal circumstances, may operate not only to destroy the tube completely but to injure other apparatus as well. Under such circumstances the overload protection for the tube bank as a whole would not be of much value. Therefore each tube is provided with a small overload relay. When the anode current exceeds a certain average value the relay contacts are opened. In this way the no-voltage release circuit of the power line breaker may be opened or the holding current of a remotely controlled switch may be interrupted, directly or thru the agency of an intermediate relay. Such a closed circuit system is preferable in this type of installation to the open-circuit type.

It will be obvious that if the flow of water in the cooling jacket is interrupted, the tube may suffer serious damage, resulting in loss of vacuum. In extreme cases the anode may be punctured. Hence, in a permanent installation some sort of alarm which operates when the flow is less than a certain amount is necessary. There are devices on the market which operate on water pressure rather than flow. These are satisfactory for certain cases, but do not afford complete security under all conditions. Two different types of alarm depending on flow have been developed. The one now in use is shown in Figure 8. Two views are presented, that at the left being an assembled alarm and that at the right being an "exploded" view. One feature of this design is the air dome formed by the cap. The imprisoned air prevents the water from reaching the electrical contacts. The advantage thus obtained in maintaining clean contacts is obvious. Such a device can readily be made to open the main circuit breaker, as well as to operate some audible or visible type of alarm.

The copper anode is safe providing that the temperature of every part is below the boiling point of water. However, to insure this condition, the temperature of the water at the outlet must be well below 100° C. This is particularly true when several water jackets are connected in parallel and only the average temperature of all is measured. No one outlet temperature can be

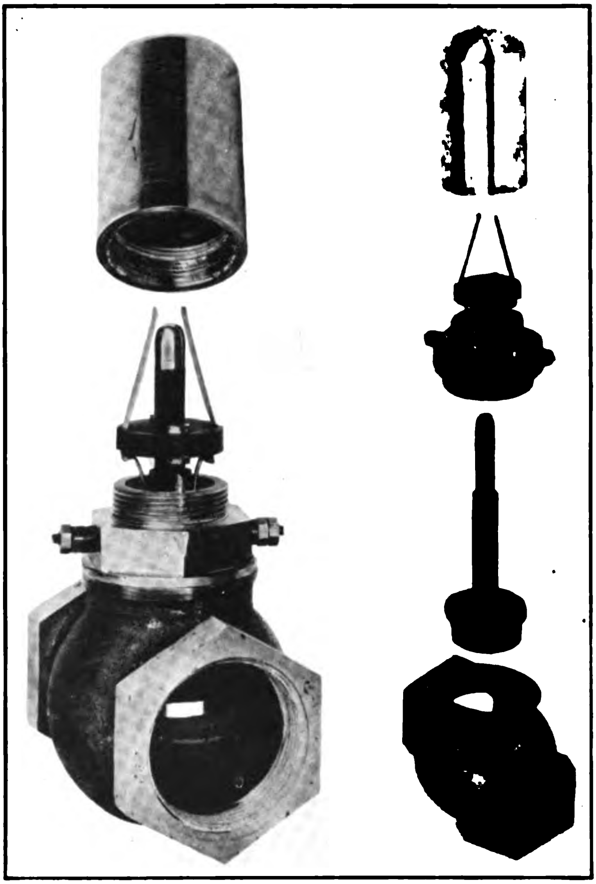


FIGURE 8—Partially Assembled and Exploded Views of Waterflow Alarm

specified for all cases since this depends on the rate of flow and the inlet temperature. It is entirely safe to operate a ten-tube unit with an outlet temperature of 60° C. when the inlet is at 25° C. and the water flow per jack is one gallon per minute.

Normally, however, the temperature is considerably less than 60° C. Means are provided to insure that this temperature is

not exceeded. A thermometer having contacts which close when a certain temperature is reached, may operate an alarm thru suitable relays or it may open the circuits which supply the anode and filament power.

INITIAL DEVELOPMENT OF A TWO-TUBE AMPLIFIER

Preliminary experiments devised primarily for the purpose of determining the utility of the present type of water-cooled tube as a power amplifier for telephone purposes were undertaken and extended into the early part of 1922.

This set will be briefly described, not only because it is used in the trans-atlantic tests, but because it possesses certain features of general interest. The assembled laboratory equipment is shown in Figure 9. From left to right may be seen the amplifier unit, the rectifier unit and the power control panel. The low power equipment is not shown. On the amplifier may be seen several meters, a thermometer for determining the outlet temperature of the cooling water, a flow indicator and relays for the individual protection of the tubes. There are also electrically-driven clocks for automatically recording the total time that filament and plate voltage have been applied.

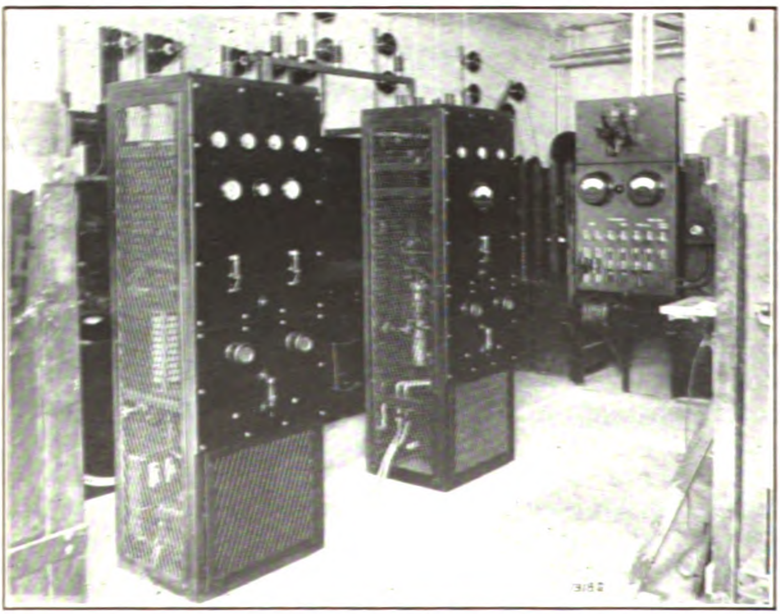


FIGURE 9—Amplifier and Rectifier Assembly
The former employs two amplifier tubes, the latter two rectifier tubes, all water-cooled

Figure 10 is a rear view of the amplifier. The space behind the panel is enclosed with expanded metal so that while the tubes are visible there is no danger that the operator will accidentally come into contact with a high potential conductor. The filament transformers, one for each tube, and such other apparatus as the potentiometer and filter used in connection with the grid polarizing circuit, are also located in this enclosed space.

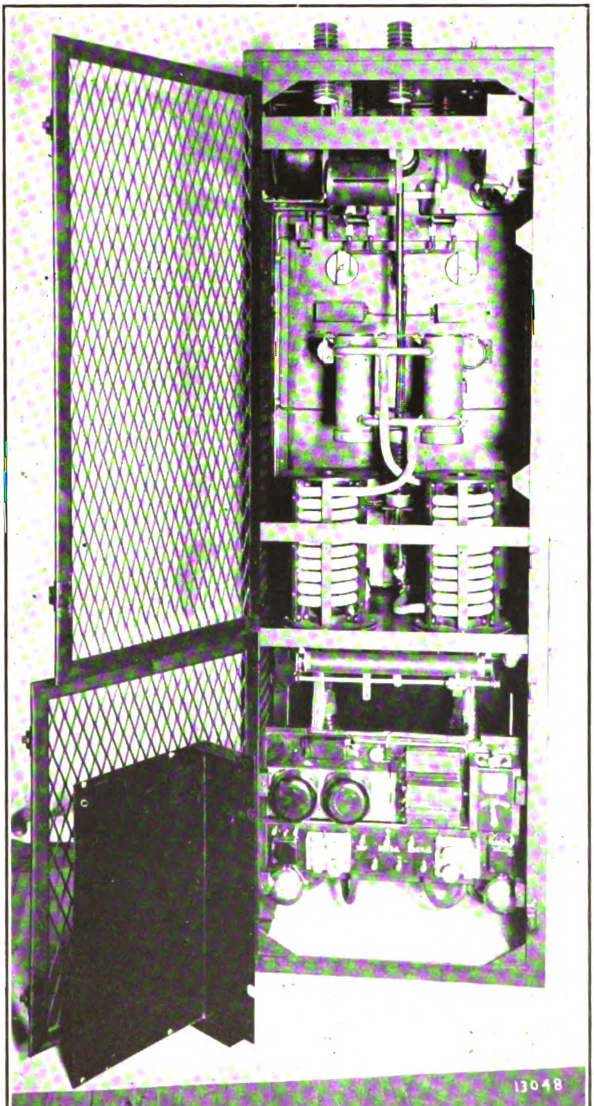


FIGURE 10—Rear View of Amplifier Set Shown in Figure 9

Just beneath the tubes are placed the insulating hose coils thru which cooling water circulates. These are necessary because the anodes operate at high voltage and the water supply is at earth potential. The general belief is that ordinary tap water, undistilled, is a fairly good conductor. Calculations made of the resistance of a half-inch hose a few feet in length filled with tap water show the resistance to be in the neighborhood of a megohm. Because of this rather high resistance these water columns need not be very long.

The rectifier was of the whole-wave type, two two-element water-cooled tubes being employed. From Figure 9 it is seen that these were mounted in a unit similar to the amplifier. A rear view is shown in Figure 11.

In a finished set the power control apparatus and the units containing the low power oscillators, amplifiers and modulators would be lodged in panels of the same height and general appearance. They would be placed adjacent to one another along the same line so that all of the routine adjustments which could be made with power on would be made from the front of the set.

The result of this development was a power amplifier unit which operated in the range of wave lengths now used for broadcasting and which could be readily converted to operate at any other suitable radio telephone frequency. Altho the power supply was a sixty-cycle single-phase rectifier, the filter circuit built to eliminate the large ripple reduced it to about one percent of the supply voltage. Had it been necessary, an increase in the size of the filter could have been made to attenuate the ripple to a lower value without making the equipment unduly bulky or costly.

Amplifier Requirements for Experimental Trans-Atlantic Installation

Since the generation of a single-sideband signal is most readily accomplished at low power levels (of the order of 500 microwatts) the complete amplifier system must receive energy at this level and increase the power to very large values. For the purposes of the present experimental installation, a maximum power output of 150 kilowatts is required from the last vacuum tube bank. This represents an over-all power amplification ratio of the order of three hundred million.

Obviously the first stages of such an amplifier system are small sensitive devices operating at low potentials and requiring careful electrical shielding. As the amplification progresses a

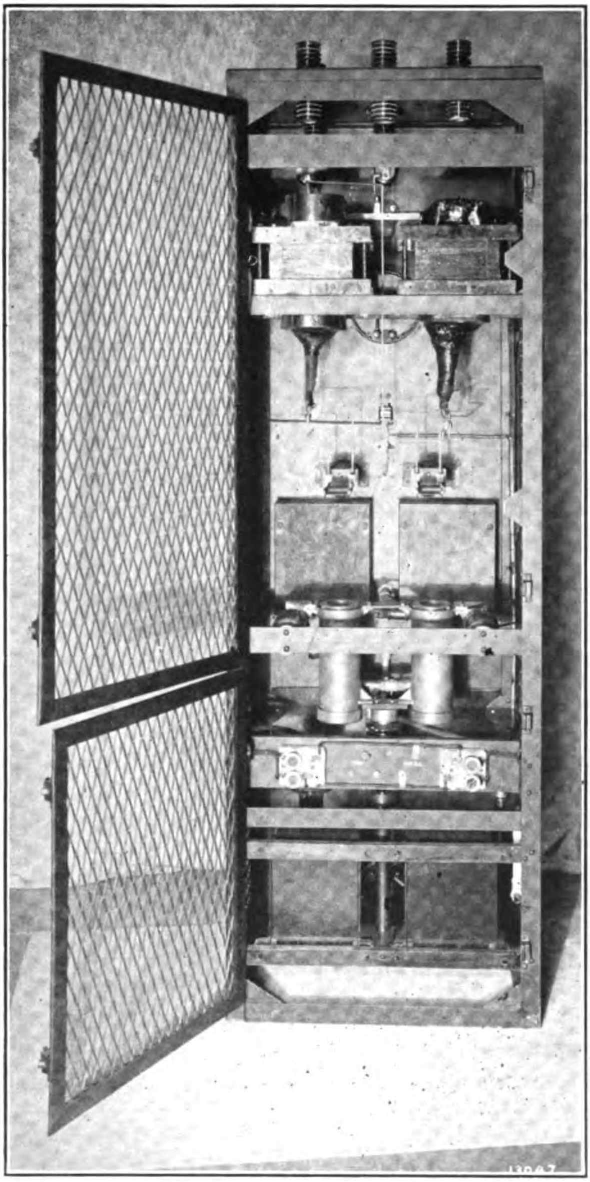


FIGURE 11—Rear View of Single Phase Whole-wave Rectifier Shown in Figure 9

point is reached where shielding becomes less important and a large upward step is taken in anode potentials of the amplifiers. The entire type of apparatus design may change at this point. Hence it is logical and convenient to divide the system into two

distinct amplifier units operating in tandem. The low voltage stages will be termed the intermediate amplifier and the high potential stages will be referred to as the power amplifier. Only the latter will be given detailed consideration.

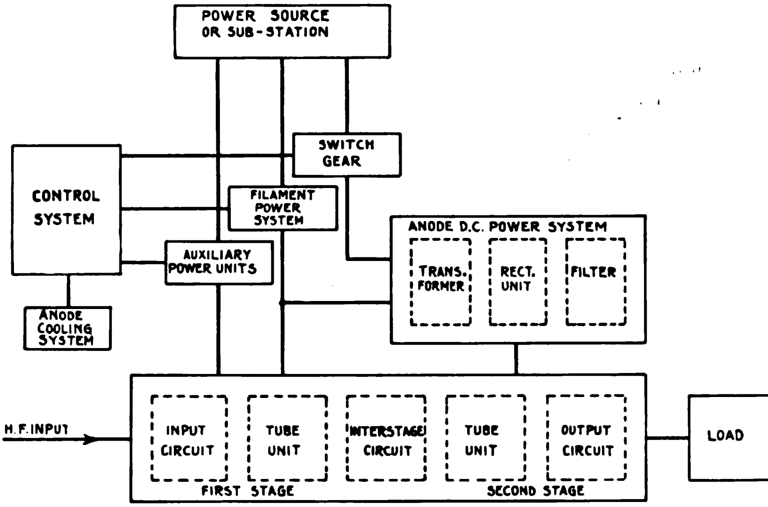
In the present installation the division between intermediate amplifier and power amplifier, expressed in terms of power level, occurs in the range 300 to 500 watts. Hence the amplification requirement of the power amplifier for the experimental system may be stated as 400 watts input and 150 kilowatts output. If there is to be no distortion the amplification ratio must be uniform thruout the transmission frequency band for all amplitudes.

An outstanding advantage of the single-sideband eliminated-carrier method of transmission lies in the fact that the frequency band over which uniform amplification is desired is reduced to slightly less than one-half of that usually required and this greatly simplifies the amplifier design problem, particularly in the case of trans-oceanic transmission where relatively low radio frequencies are employed. This matter will again be considered when discussing the amplifier circuit design.

Block Schematic of Power Amplifier System

Due to the large power capacity and the relatively high operating potentials, the design and construction of the experimental power amplifier system involved a number of new problems. The size and weight of various parts precluded the ordinary procedure of mounting the vacuum tubes and the associated apparatus in a single self-contained unit. The arrangement adopted was not unlike that for other types of power apparatus where the system comprises a number of complete units, separately located and controlled from a central position. Altho, in the case of an amplifier system the design of each unit is closely associated with the others, it will be convenient to consider them independently and to subdivide and restate the problem as applied to each part. For this purpose the block diagram shown in Figure 12 has been prepared.

The general problem of the power amplifier system has already been stated; viz., to receive a single-sideband signal of approximately 400 watts maximum power, distributed over a frequency band of 2,500 cycles, in the frequency range of 55,000 cycles to 60,000 cycles, to amplify this signal with minimum distortion to 150 kilowatts maximum power and to deliver the amplified signal to an antenna system. This requires a power amplification ratio of 375.



BLOCK DIAGRAM
TWO STAGE POWER AMPLIFIER
FIGURE 12

When dealing with power outputs of 150 kilowatts, operating efficiency is an important item. High efficiency is obtained without undue distortion by employing amplifiers of the third class, but since the grids of such amplifiers are driven positive, an appreciable power is expended in the grid circuit of each stage, thereby limiting the amplification per stage to ratios much less than those secured with amplifiers of the first class. In power amplifiers of the third class a ratio of 375 requires two stages. It may be possible to obtain the same amplification at these levels by means of a single stage, but more tubes will be required.

In Figure 12 these two stages are divided into five units. The tube units comprise mountings for the tubes in each stage, means for supplying cooling water and filament heating current, and such individual protection as each tube may require. They do not contain the radio frequency power circuits common to all the tubes in a unit. These radio circuits are designated as the input circuit, the interstage circuit, and the output circuit.

The input circuit serves to connect the grid-filament terminals of the first tube unit with the line carrying the signal which is to be amplified. One of the functions of this apparatus is to receive efficiently the incoming signal energy and to convert it to the proper potential for operation of the power tubes; that is, to match the effective grid-filament impedance with that of the line

thruout the signal frequency transmission band. Another function is to introduce the desired grid polarizing potential and to provide amplification control. The interstage circuit serves a similar purpose between the anodes of the first tubes and the grids of the second.

The output circuit connects the anodes of the second tube unit with the load. It acts as a conversion circuit which efficiently transfers the radio frequency power to the load circuit, which is, of course, the antenna system. The output circuit is required to eliminate radio frequency harmonics, and in cases where the impedance characteristic of the antenna is unsatisfactory, the output circuit must be designed to correct the difficulty. This last requirement cannot always be completely met, but some correction is usually possible, as will be shown later.

The anode direct current power system must convert the available supply into direct current power at the proper voltage; it must suppress noise from the power source, and it must provide a path to by-pass the large voice frequency components generated in the anode circuits of the amplifier.

The functions of the remaining blocks shown in the diagram of Figure 12 are self-explanatory. Their requirements are largely detail matters relating to control and protection, both for the power units and the tubes. Since the tube circuits are supplied with power thru three lines and are dependent upon the proper circulation of cooling water, it is apparent that the control circuits must possess interlocking features, which prevent incorrect applications of power when starting the plant and which avoid damage to the system by switching off power whenever an abnormal condition is established during operation. Further than this, the control system should signal the attendant and give some indication of the kind and location of trouble. It should also include suitable safety devices designed to prevent the attendant from accidentally examining high potential parts without shutting off the power. All of these things can usually be accomplished in a number of different ways and the exact arrangement is determined by the particular conditions.

150-Kilowatt Amplifier for Trans-Atlantic Experiments

It has been shown that a two-stage system is required to amplify a 400-watt signal to a 150-kilowatt signal efficiently and economically. Hence the experimental system was constructed on the basis of two 10-kilowatt tubes for the first stage and twenty 10-kilowatt tubes for the second. Since a two-tube unit

is capable of delivering 20 kilowatts, there is an ample margin in power capacity, and advantage is taken of this to adjust the circuit for improved efficiency and better quality. The reasons for using twenty instead of fifteen tubes in the last stage will be more apparent later when the design of the output circuit is considered. It will be sufficient at this point to say that the load power factor at the edges of the telephone transmission band may depart considerably from unity and that this effect, in combination with the anode circuit impedance characteristic, necessitates the extra volt-ampere capacity.

The two-tube unit which was developed shortly after the first experiments with the present water-cooled tube and which has already been described was available for the first stage of the power amplifier. The single-phase rectifier unit was not required. The tube panel as installed for the first stage is shown at the right in Figure 13. In the middle of the picture is shown the central control panel for the power amplifier. This panel does not include remote control of the radio frequency tuning apparatus because the cost of such control is not warranted in an experimental plant.

The twenty power tubes for the second stage were arranged in two panel type units each containing 10 tubes. The object in constructing two separate tube units was to provide greater flexibility both for experimental and routine testings. There are several operating advantages to be gained by dividing the tubes into two groups such, for example, as running at reduced power with one tube unit, while repairs are being made to the other unit. The suppression of undesired oscillations becomes more difficult as the number of tubes in a single bank is increased. During the course of the initial testing of the 150-kilowatt stage there were numerous occasions when the division of the tubes into two banks proved to be a valuable aid.

The two ten-tube units for the second stage of the power amplifier are shown at the left of Figure 14. Mounted on the upper panel are two auxiliary control buttons for use in emergencies, and a thermometer with electric contacts to ring an alarm in case the cooling water temperature exceeds a safe limit. The plate glass panel gives the attendant a full view of all the tubes and of the scale of the ammeter used for checking individual anode currents. This meter is located at the forward end of the high voltage rack about seven inches behind the glass panel. The meter is switched into and out of each anode circuit by means of jacks operated by a metal plug on a long insulated rod. The

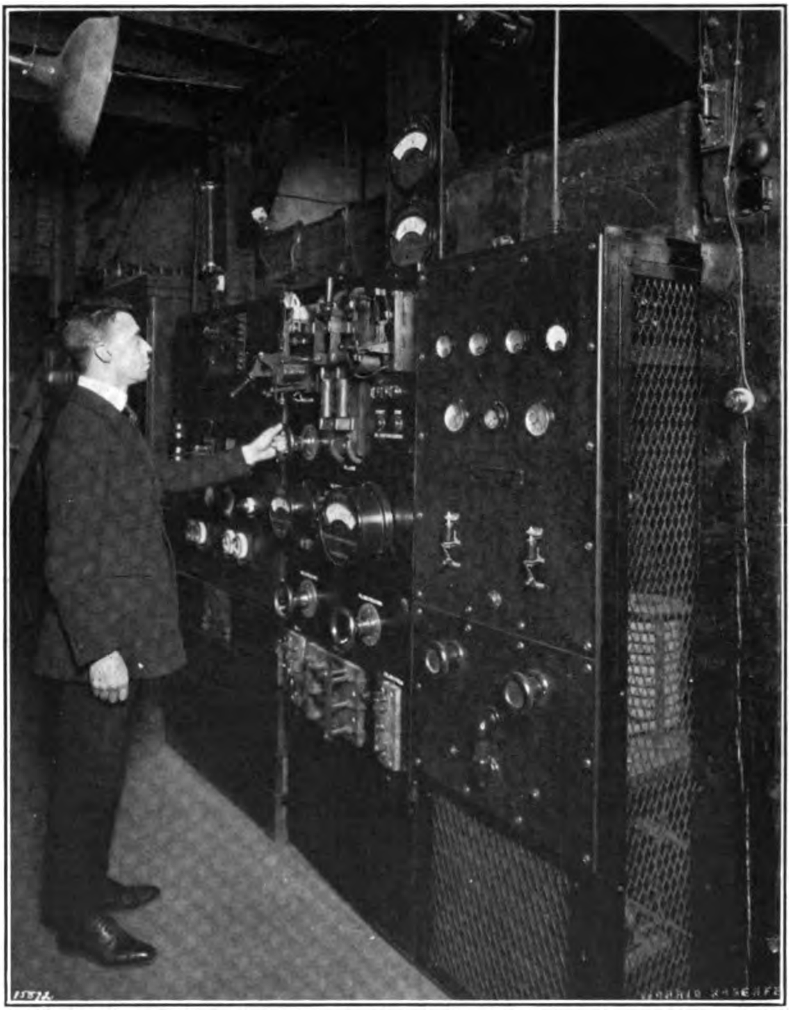


Figure 13—Amplifier Shown in Figure 10 Installed as the First Stage of the Power Amplifier. The Second Stage is Shown in Figure 14

two lower panels carry the rheostats and switches for individual control of filament currents.

The jackets for the ten-tube unit are arranged in two rows of five tubes each as shown in Figure 15 and Figure 16. Compact arrangement of tubes is highly desirable both from an electrical and mechanical viewpoint. Cooling water is circulated thru common inlet and outlet headers so arranged that the flow is equalized for all tubes. The connection between each jacket and the header contains a very short length of rubber hose for the

purpose of providing a high resistance electrical path between jackets. The reason for this will be evident when the intertube electrical circuit is described. No valves are provided to shut off the flow to individual jackets. In case it is desired to operate without a complete set of tubes, metal disc stoppers are substituted for tubes. It is not necessary for this stopper to have the same shape as the anode because the pressure drop thru the jacket is small compared to that in the connections to the headers.

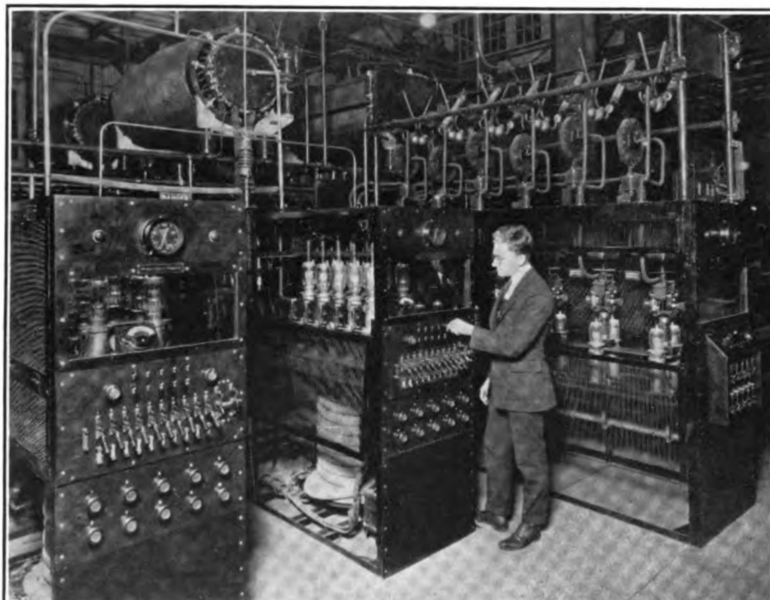


FIGURE 14—Last Stage of Amplification. Two ten-tube sets are shown, also the six-phase rectifier at the right

Since the cooling jackets and the common intake and outlet headers are at the same potential as the anodes, it is necessary to bring the water to and from headers by means of hose coils as previously explained when describing the two-tube unit. These hose coils are located directly under the insulated frame supporting the water jackets. The intake and outlet lines are wound in parallel on the same frame. One coil is clearly visible in Figure 14.

The water jackets are insulated from each other for comparatively small voltages and are mounted on a main frame which is insulated for the full anode potential. This frame also supports the radio frequency bus, the individual by-pass condensers,

FIGURE 15—Front View of One of the Amplifier Units

the individual anode circuit overload trip-coils, the network for the suppression of inter-tube oscillations and the meter, jacks and cables necessary for checking the individual anode currents.

The grids are connected directly to a common grid bus as shown in Figure 16. One side of the filament is connected in like manner. The other side of each filament is connected to a rheostat and switch controlled from the front of the panel unit. Filament heating current is supplied to the ten tubes by a transformer placed on the floor between the hose coil and lower panel board.

Adequate protection against high potentials is provided by expanded metal screens arranged for quick removal to facilitate repair work. These screens also introduce a certain degree of electrical shielding between the banks. Altho such shielding does not appear to be essential, its presence undoubtedly helps to stabilize the anode-filament capacity relations.

A schematic diagram of the electrical circuit comprising one ten-tube unit of the second power amplifier stage is shown in Figure 17. It will be noted from this that the problem of connecting ten power tubes for parallel operation involves considerably more than merely connecting the terminals to a common bus system.

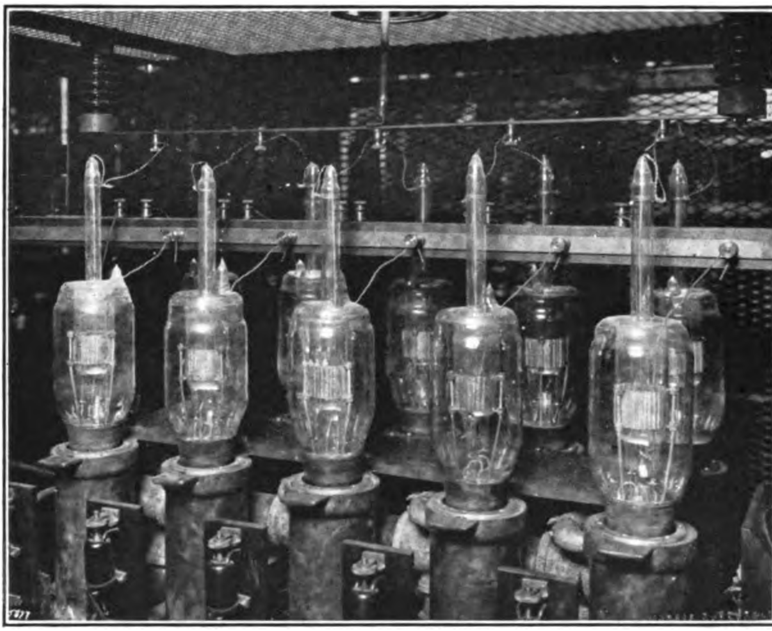


FIGURE 16—Side View of One of the Amplifier Units, Showing Tube Arrangement

Beginning with the filament circuit, there are two buses connected to the terminals of the heating transformer secondary. These are provided with the usual bypass condensers. The direct current from the anode circuit passes thru the transformer winding to ground. In order to balance the direct current thru the two halves of the transformer winding it is necessary to place five filament rheostats in the lead to one bus and the remaining five in the lead to the opposite bus.

It will be recalled that in describing the arrangement of water jackets and inter-connecting water lines it was stated that the jackets were insulated from each other. The jacket, of course, corresponds to the anode in Figure 17. In order to eliminate certain types of inter-tube oscillation, all of the anodes are connected together thru a resistance network so designed that sub-

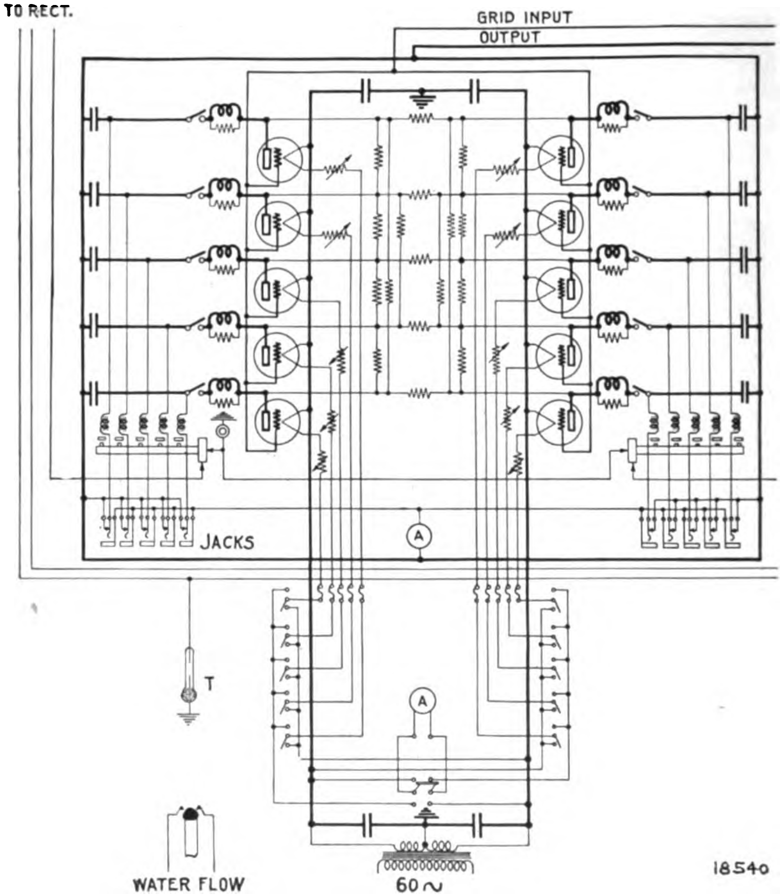


FIGURE 17—Schematic of the Electrical Circuit of One Ten-tube Amplifier Unit

stantially equal resistance paths exist between any two jackets. Each anode is connected to the main bus thru a path including a resistance and a choke coil in parallel. The latter passes the normal radio frequency. The underlying principle of the network is to provide a low impedance path for the normal radio signal and a non-oscillatory resistance path for undesired oscillation. The inductances lower the resonant frequencies of the circuits involved in the tube-to-tube oscillations to such an extent that the grid coupling, which is due to the inductance of the grid leads, is insufficient to sustain oscillations when damped by the resistance as described. Such a network is equally effective when placed in the grid circuit insofar as suppression of undesired

oscillation is concerned, but there are other reasons for not doing this. For example, since the grid is driven positive during a large part of each cycle, there is an appreciable grid current and every impedance in the path of this current tends to increase the non-linearity of the input-response characteristic of the amplifier. Then again, obstructions in the grid circuit will presumably tend to aggravate any effects due to secondary emission or to unusual gas conditions. The latter occurs only in exceptional cases when a tube develops a slow leak, but the consequences may prove very injurious to the adjoining good tubes.

The direct current component flowing to the anode passes thru overload trip coils associated with an auxiliary control circuit. Experience thus far has shown that such devices give valuable protection. The jacks connected in series with these trip coils are so arranged that the insertion of a cordless plug switches an ammeter into the plate circuit. The information concerning tube performance obtained by observing the individual anode currents more than justifies this slight switching arrangement.

A simplified overall schematic of the power amplifier system is given in Figure 18. The input circuit to the first stage is simple because the output impedance of the intermediate amplifier, which supplies the signal at the 400-watt level, is approximately equal to the correct input impedance for the grids of the two-tube unit. Under these circumstances direct resistance coupling is satisfactory. A slightly greater output could be obtained from the first power stage by driving the grids thru an input transformer, but the additional power is unnecessary. The resistance coupling arrangement has a very desirable transmission characteristic and requires no difficult adjustments. A large inductance connected in parallel with the resistance coupling element bypasses the direct current component of the grid current. The polarizing potential is supplied from a generator connected thru a potentiometer and a filter.

The interstage circuit consists of a parallel tuned circuit designed to match the impedance of the two power amplifier stages. The resistance element flattens the impedance characteristic, thus making the circuit suitable for the transmission of telephone signals, and it also assists in preventing the generation of spurious oscillations at frequencies of the same order as those of the signal. The grid polarizing potential is supplied thru a filter system the impedance of which is made low to prevent a blocking action which may occur under certain conditions. The

direct current to the grid is from one to three amperes and, in order to avoid distortion caused by changes in the grid polarizing potential which result from variations of this current, the impedance of the potentiometer from which the polarizing potential is taken must be relatively low. For this reason about 2.5 kilowatts are expended in the potentiometer.

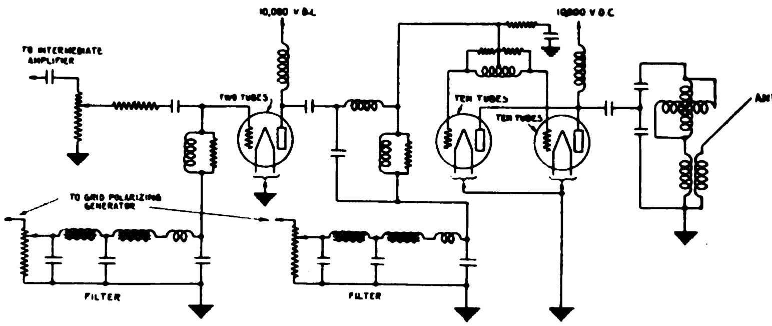


FIGURE 18—Simplified Over-all Schematic of the Power Amplifier System

The grid buses of the two 10-tube units are connected to the common grid circuit thru a mid-tapped reactance coil having a high mutual coupling between the two halves. The signal frequency currents pass thru this coil in opposite directions and are impeded only as a result of the leakage reactance. A resistance is connected in parallel with the coil. Any disturbances tending to establish oscillations between the two tube-banks encounter the full impedance of the coil with its shunted resistor. In doing this they are suppressed because of the large resistance component of this impedance. The same device can be applied in the anode circuit. In this case the coil must have a very much larger current carrying capacity.

The Output Circuit and Load

In subdividing the amplifier problem with reference to Figure 12 the output and load circuits were defined. It will be recalled that one of the functions of the output circuit is to match the impedances of the power tubes and of the load. That is, having given an amplifier capable of delivering a certain amount of radio frequency power and an antenna with certain characteristics, it is necessary to design an intermediate circuit which satisfies the impedance requirements both of the power tubes and of the antenna. A second condition which the output current must satisfy is that the power at all frequencies in the band must receive sub-

stantially the same amplification in passing thru the system to the antenna. In addition, in order to prevent interference, currents of harmonic frequencies must not be allowed to flow in the antenna.

From the foregoing requirements it is obvious that in order to design the output circuit it is necessary to know something about the characteristics of the antenna. One of the multiple tuned antennas at the Radio Central Station of the Radio Corporation of America was used. It is located at Rocky Point, Long Island. For a discussion of multiple tuned antennas the reader is referred to articles by E. F. W. Alexanderson and E. E. Bucher, in the "General Electric Review" for October, 1920, and to one by Alexanderson, Reoch, and Taylor in the American Institute of Electrical Engineers "Proceedings" for July, 1923.

The resistance component of the antenna impedance as viewed thru the feed current downlead, that is, the series resistance of the antenna, is a function of the loading in the various downleads. An experimental curve is shown in Figure 19. As

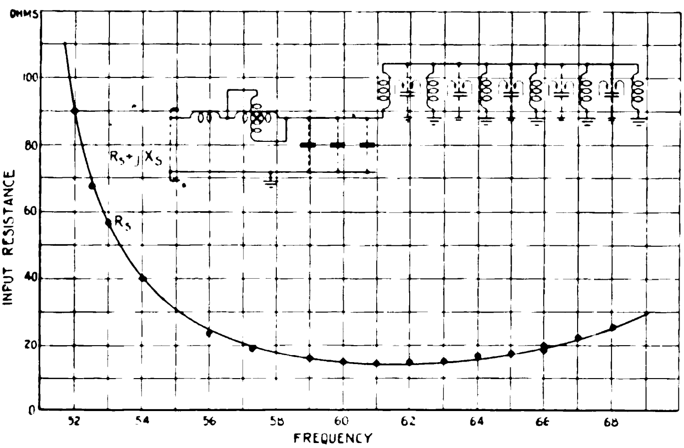


FIGURE 19—Resistance Characteristic of Multiple Tuned Antenna

the frequency increases from the lowest shown in this curve the resistance decreases sharply at first, passes thru a flat minimum and begins to rise again. At the frequency used (57 kc.) the multiple resistance of the antenna is approximately 0.76 ohms. The radiation resistance, taking account of the directive effect, is 0.32.[5] The radiation efficiency is therefore about forty percent.

[5] The effective height as determined from field strength measurements is 85 meters. The value obtained when the directive effect is neglected is 0.41 ohms. The radiation efficiency would thus be nearly fifty percent. As a result of the directive effect there is a saving in total antenna power of about 12 percent for equal signals along the axis of transmission.

This low antenna resistance is of itself most desirable, but it results in a narrow resonance curve which for telephony leads to certain complications not encountered in the telegraph service for which the antenna was designed. When the antenna is tuned to 57 kc. the band width is 1,150 cycles. (See Figure 20.) The band width is here arbitrarily taken as the difference between the two frequencies for which the impedance of the tuned antenna circuit is double the minimum value. It will be seen that the transmission of high quality speech under such conditions presents a difficult problem in spite of the fact that the system is favored by the single-sideband method.

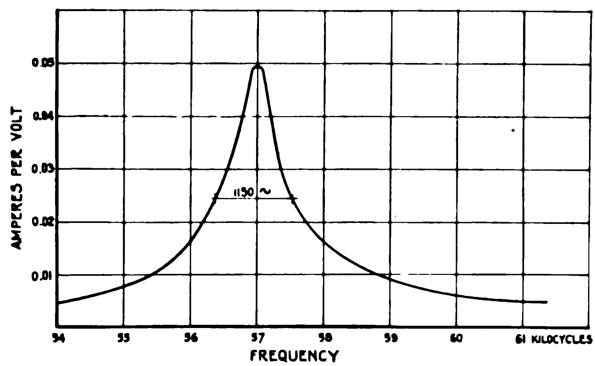


FIGURE 20—Antenna Input Admittance of the Antenna when Tuned to 57 kc.

Figure 20 does not indicate the total antenna current but only that in the downlead thru which power is supplied to the antenna. The ratio of the former to the latter is shown in Figure 21, curve *A*. The ordinate gives the sum of the six downlead currents divided by the feed current as measured in the station at *M* in the diagram. Between the station and the first loading coil there is a lead the capacity of which to earth, plus that of the coil itself, is of the order of 0.001 microfarad. Hence the current in the downlead itself is less than that in the station by an amount which is by no means negligible. Curve *B* gives the feed ratio referred to the feed current in the downlead beyond the loading coil at *N*. The difference in the present case is large because the loading inductance at *M* is a large fraction of the total loading of the first downlead. In Figure 22 are plotted the currents in the downleads divided by that at *N*. Their intersection at a frequency of about 58 kc. is the result of a proper choice of the inductances of the various downleads. The divergence above and below this point is due to the inductance of the flat-top.

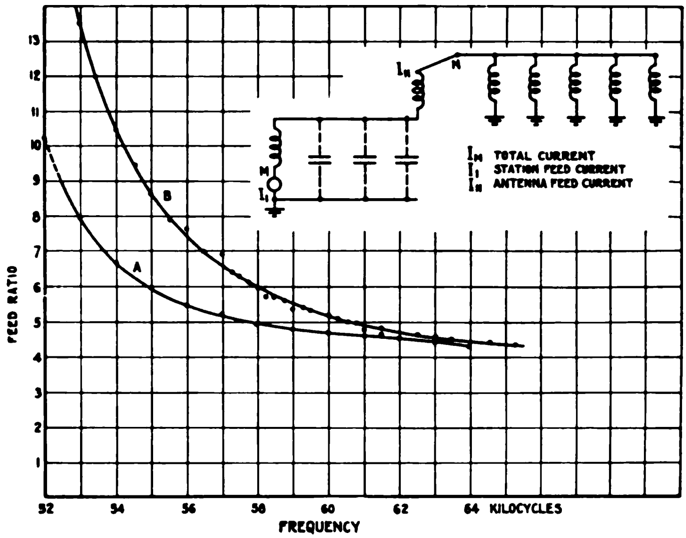


FIGURE 21—Feed Ratio Characteristic of the Antenna. The ordinates of A are the ratios of I_M to I_1. Those of B are the ratios of I_M to I_{11}

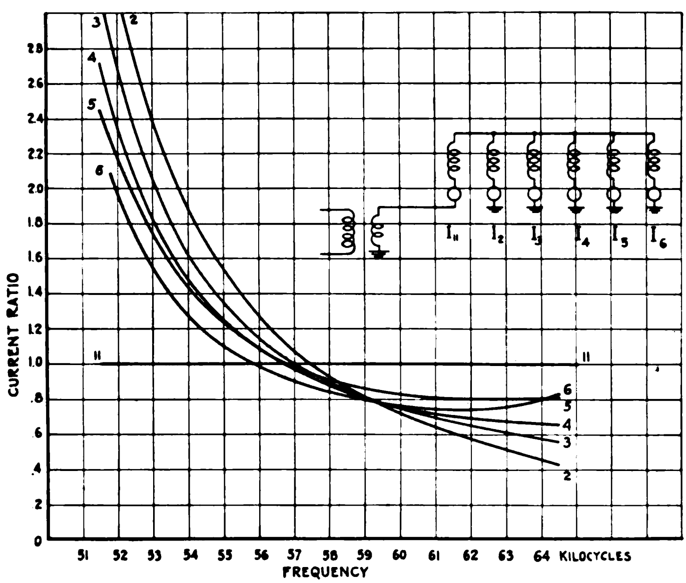


FIGURE 22—Ratio of Currents in the Various Downleads to I_{11}

There are several forms in which the output circuit may be arranged. In addition to the conditions previously mentioned, convenience, ease of adjustment, and availability of apparatus

should be considered in the design. A schematic of the circuit employed is shown in Figure 18. It consists of a capacity branch which provides a low impedance path for harmonics and an inductive branch, coupled inductively with the antenna circuit. The latter is tuned at 57 kc. The primary circuit is usually adjusted so that at 57 kc. the phase angle of the load into which the amplifier works is approximately zero; that is, its power factor is unity. All of the adjustments necessary in routine operation can be made by means of variometers and a variable coupling coil.

In a single-sideband eliminated-carrier system this tuning frequency of 57 kc. is an upper sideband frequency corresponding to a 1,500-cycle audio signal applied to a 55.5 kc. carrier. Since 1,500 cycles is approximately at the middle of the voice frequency band which it is desired to transmit, it follows that the antenna and output circuits when tuned to 57 kc. are in correct adjustment for the middle of the upper radio frequency sideband. Hence the initial electrical design considerations are centered about this frequency.

Having fixed upon the type of output circuit, the electrical design procedure consists of two steps: first, the preliminary determination of circuit constants on the basis of a single tuning frequency as stated above and, second, an examination of the frequency impedance characteristic thruout the band for the purpose of modifying the constants so as to obtain the most practical input-response-frequency characteristic for the amplifier system.

Referring to Figure 23, let the antenna reactance, X_s, be

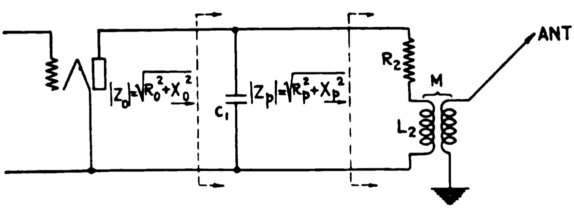


FIGURE 23

zero for the frequency f_t. Let R_s equal the corresponding antenna resistance and R_{s_2} equal the resistance component introduced into the inductive branch of the output circuit thru the coupling M. Let R_2 represent the total resistance of the output circuit alone and assume that it is lumped in the inductive leg. This assumption introduces very little error because

the coil losses are usually several times greater than the condenser losses. The total resistance of the inductive leg is therefore

$$R_p = R_2 + R_{s2}$$

Since the antenna reactance is zero, the reactance component, Xs_2, introduced into the inductive branch of the output circuit is also zero and X_p is equal to X_2, where X_2 is the reactance of the inductive branch alone. In the simple case shown in Figure 23, X_2 is equal to $2\pi f_t L_2$. In general the impedance, Z_o, of the output circuit as seen by the tubes has both a resistance component R_o and a reactive component X_o. It is apparent that if the tubes are to work into a pure resistance, X_o must equal zero for the frequency f_t, in which case $Z_o = R_e$ where R_e is the particular value of R_o corresponding to the frequency f_t. The conditions for unity power factor are satisfied when

$$R_p = \frac{X_1^2 R_e}{X_1^2 + R_e^2} \tag{1}$$

and

$$X_2 = \frac{-X_1 R_e^2}{X_1^2 + R_e^2} \tag{2}$$

where X_1 is the reactance of the capacitive branch of the output circuit. The design is started by fixing the numerical value of R_e. The correct determination of R_e requires previous experimental knowledge concerning the type of tube to be employed because it involves the effective internal tube resistance and the latter in turn is a function of the anode mode of operation. If N is the number of tubes connected in parallel, then $R_e = \frac{K}{N}$ where K is constant for a given application. When water-cooled tubes of the type already described are supplied with direct current at 10,000 volts and are operated as single frequency amplifiers of the third class, 5,000 ohms is a good value for K. If transmission occurs over a band of frequencies, then the choice of the value of K at the middle of the band will depend upon the way in which the impedance varies for higher and lower frequencies. If the output circuit is to prevent anode circuit harmonics from reaching the antenna, such currents must pass thru condenser C_1. The impedance of C_1 to the second harmonic should be of the order of one-fourth of the impedance offered by the output circuit to the fundamental. That is

$$C_1 = \frac{1}{\pi f_t R_e} \tag{3}$$

[illegible]

$$[illegible]$$

[illegible]

$$[illegible]$$

[illegible]

$$[illegible]$$

[illegible]

$$[illegible]$$

[illegible]

[illegible] This [illegible] The [illegible]

[illegible] further [illegible] antenna [illegible] For lower [illegible] the fre- [illegible]

maximum. For frequencies higher than 57 kc., capacity reactance is added to the inductive leg of the output circuit, making the latter anti-resonant at the frequency of the peak, C. The impedance at these peaks is high because the resistance introduced into the primary from the antenna both above and below 57 kc. is much lower than at the tuning frequency.

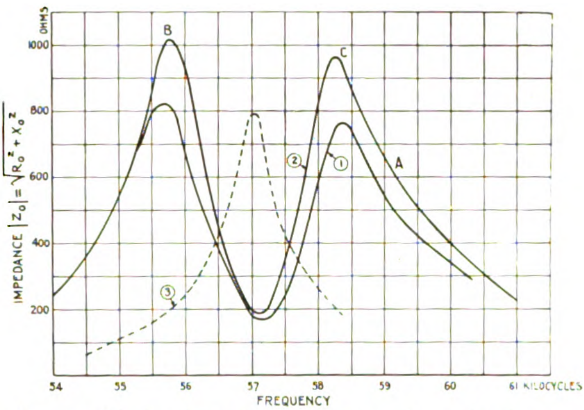


FIGURE 24—1. Impedance of the output circuit as a function of frequency, measured values. 2. The same, calculated values. 3. Resonance curve of antenna

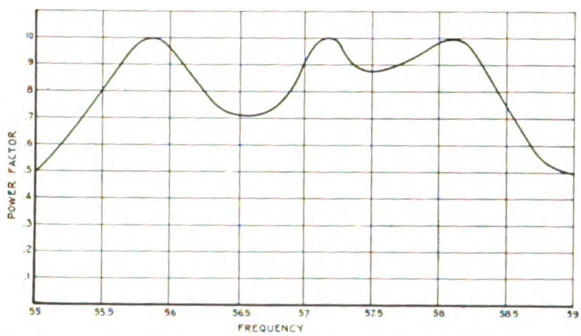


FIGURE 25—Power Factor of the Output Circuit. The middle maximum corresponds to the tuning frequency, the other two to points B and C of Figure 24

Let us now see how this is related to the antenna current-frequency characteristic, assuming that we apply a constant voltage of varying frequency to the grid of the last stage. If the impedance of the tubes were zero, this would be equivalent to the application of a constant voltage equal to μe_c (amplification constant times alternating grid voltage) to the output circuit. The antenna

[illegible]

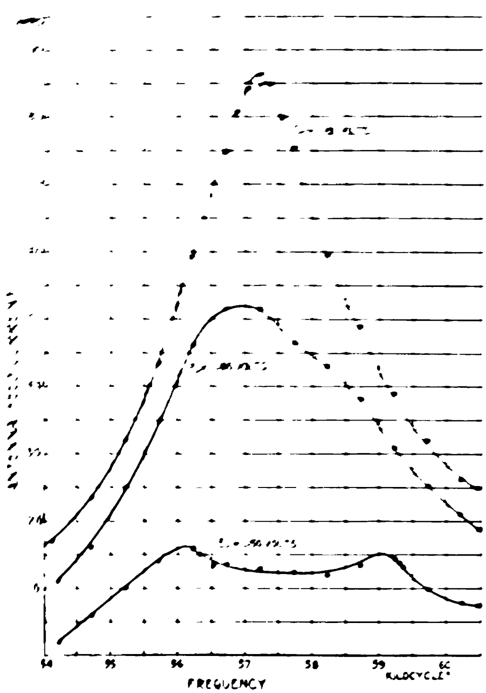


FIGURE 26. Effect of Tube Impedance on the Transmission Characteristic at Low Amplitudes

three curves for different values of grid polarizing voltage are shown. The upper one is for $E_c = -50$ volts. The tube impedance is therefore lower than usual. The middle and lowest curve corresponds to -100 and -150 volts, respectively. The forma-

tion of a second maximum at 59 kc. may be traced. Even in the upper curve this tendency can be seen, but not until the tube impedance is that corresponding to the lowest curve does the new maximum become equal to the first. It will also be noted that the position of the original peak has apparently been shifted to the left.

The characteristics which have just been discussed were taken at low amplitude such that the operation of the amplifier was probably as one of the first class. When the grid input is so large that the grid during a portion of the cycle assumes high positive values and the highest power output is obtained, the characteristic corresponding to the low amplitude curves does not in general have the same shape. This is due to the fact that under such conditions the tube impedance is not constant but depends on the alternating grid and plate potentials as well as upon their relative phase. This is illustrated by Figure 27 in which Curve 1 is for low amplitude and Curve 2 for high.[6] In curve 1 there are tendencies to form maxima at 56 and at 58.5 kilocycles. In passing to Curve 2 the tube impedance at these frequencies becomes much greater than at 57.4 kilocycles because of the tendency for the anode to assume small or negative values when the grid potential is positive.[7]

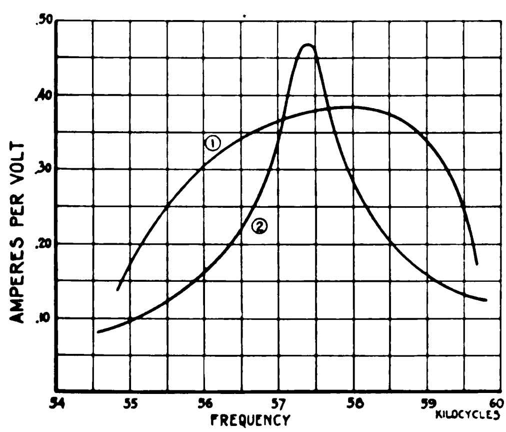


FIGURE 27—Transmission Characteristics. 1. Low Amplitude. 2. High Amplitude

[6] This figure should not be compared quantitatively with others in this paper since it is for a different circuit adjustment.

[7] In speaking of the "impedance" of a tube operating as a third class amplifier, we are thinking in qualitative terms only. Tube impedance does, of course, have a real and useful significance for an amplifier of the first class, but for the third class great care must be taken lest we beguile ourselves into thinking that we are talking about a constant. This mistake, however, is often made.

In Figure 28 is plotted the antenna current as a function of grid input, the frequency being constant at 57.2 kc. Altho this is not a straight line, as is desired for perfect reproduction, it is nevertheless a fair approximation to this up to the knee of the curve. This will be seen by comparison with the dotted line. The problem of obtaining a straight amplitude characteristic is one which is not met in building an ordinary telegraph transmitter. This is because only two amplitudes are of interest, viz., maximum and zero,[8] Hence, it is somewhat more difficult in telephony to obtain the best efficiency because the grid polarizing potential must be less than the value which, in the absence of alternating grid input, reduces the anode current to zero.

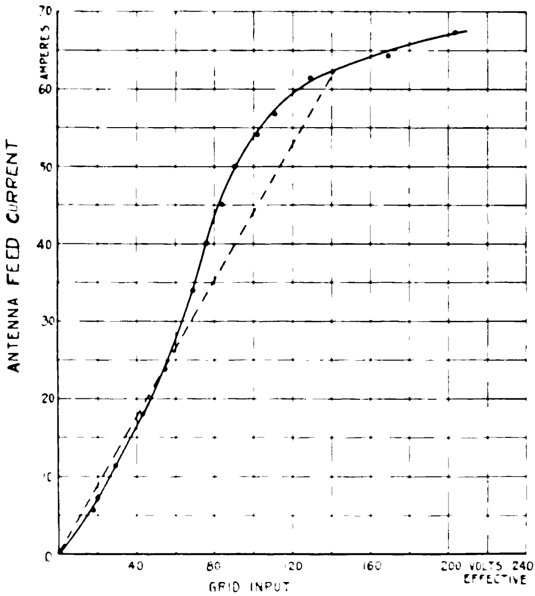


FIGURE 28—Input-response Curve at Constant Frequency

It has been mentioned that in an output circuit such as has been described, the impedance at points in the transmission band may be very high. This fact results in one advantage and one disadvantage. The advantage is the widening of the input-response-frequency characteristic. This has been discussed above. The disadvantage results from the fact that the widening is accomplished only at the expense of a diminished factor of

[8] This statement should not be taken to include a telegraph repeating system which receives and retransmits a signal without relays.

safety. For with increase of output impedance there is a corresponding, tho not proportional, increase in the alternating component of the anode potential. The result is that during a portion of the cycle the anode becomes negative by a value which may be as much as one-half the voltage of the direct current supply. There is no danger in this, at least under ordinary circumstances; but in the other half of the cycle, since the wave is symmetrical, the potential is of the order of 2½ times the supply voltage. In fact, there have been cases in which even greater potentials have been measured with a peak voltmeter. Hence, if the same factor of safety is to be maintained, a lower supply voltage must be employed and a decreased tube power rating therefore must follow. Thus in Figure 29 a case is illustrated in which this phenomenon takes place. Note in particular that at a frequency of 59.5 kc. the alternating peak is 15 percent more than the supply voltage. This difference may be increased by augmenting the grid input voltage.

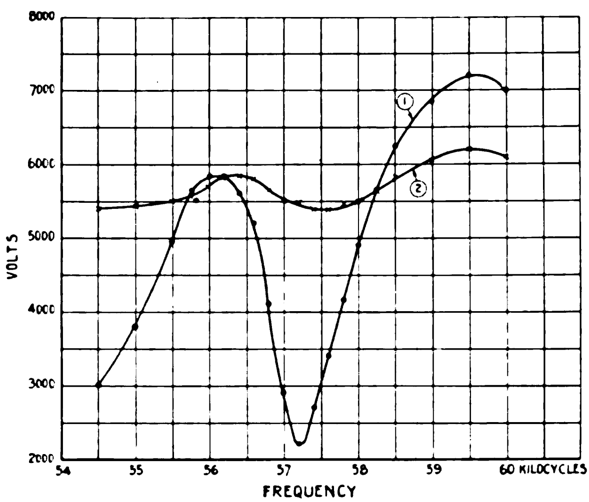


FIGURE 29—Curves Illustrating the Variation in Alternating Anode Voltage When Working Into an Output Impedance Having a Characteristic Similar to That Shown in Figure 24. Constant Grid Input Voltage. 1. Alternating peaks. 2. Supply voltage as read on direct current voltmeter

A further disadvantage in permitting the anode potential to assume negative values during a part of the cycle is that this is usually accompanied by a sudden increase of harmonics. The reason, of course, is that in this interval there is no anode current, and the abrupt changes in current at the beginning and end of the period result in strong harmonics of a fairly high order.

The power supply at Radio Central Station is three-phase sixty cycles at 22,000 volts. A six-phase rectifier was employed. The circuit is shown in Figure 30. It will be seen that the six

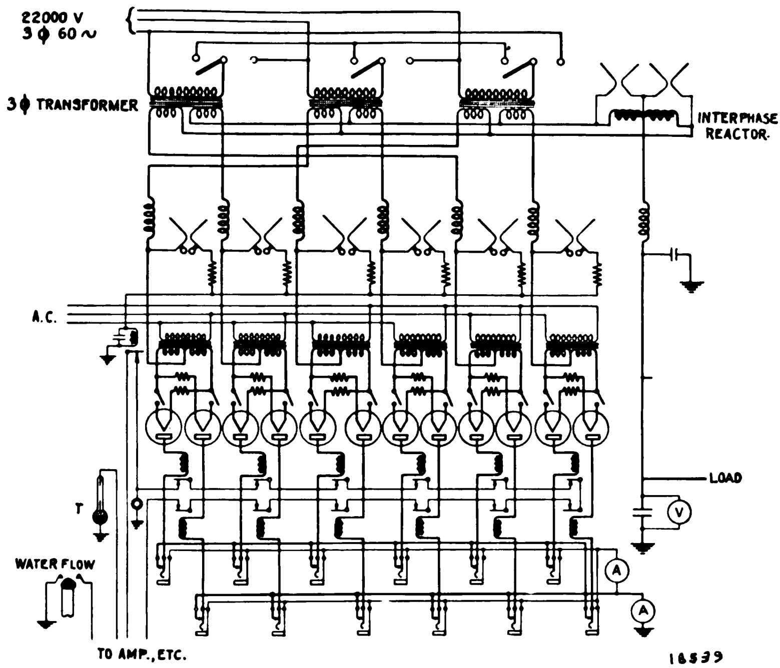


FIGURE 30—Schematic of Six-phase Rectifier Circuit

phases are connected in two groups, each a three-phase star. The neutrals of these stars are joined by an interphase transformer or reactor having a center tap from which the load is taken. The potential from these two neutrals to earth, besides having a direct voltage component, contains frequencies of 180 cycles (the third harmonic of 60 cycles) and integral multiples of 180. It is easily seen that in a symmetrical system the corresponding components of the two stars are equal for the even multiples and opposite for the odd. Since the voltage of the mid-point of the reactor is the average of those of the two ends, the components 180, 540, and so on, do not appear in the output while the 360, 720, and so on, are present with values the same as in each star. The resulting percentage ripple is quantitatively the same as that to be expected in a simple six-phase system. The even harmonics of 180 cycles can be reduced by the use of an inductance in series with the load, as shown in the diagram.

The advantage of the interphase reactor is therefore not to

be sought in any marked effect on the ripple. Its usefulness results from its action in lengthening the time when current flows thru the tubes. Among the advantages gained by its incorporation in a six-phase system are the lower thermionic emission required for a given output, less power dissipation in the anodes and better regulation. There are certain disadvantages, but they will not be discussed now.

If the rectified output from a group of phases is made constant, as for example by suitable chokes, and if during a part of the cycle one of the phases carries the whole load, it is well known that the wave form of each phase has a flat top. The statement that the wave form is rectangular is, however, incorrect. As is to be expected, it more closely resembles a trapezoid. In the oscillograms of Figure 31, we have the actual wave form of the potential of the filaments with respect to the anodes for a pair of rectifiers connected directly in parallel forming one of the six phases of the system shown in Figure 30. Below it is the approximately trapezoidal current through these tubes. From the oscillogram it may be found that the maximum current is slightly more than three times the average, as we should expect. The flow of current lasts 0.46 cycles instead of 0.33 cycles as it would were the transformer and tube impedances negligible. The maximum voltage across the tube is about 2.1 times the average, which also is in accordance with theory.

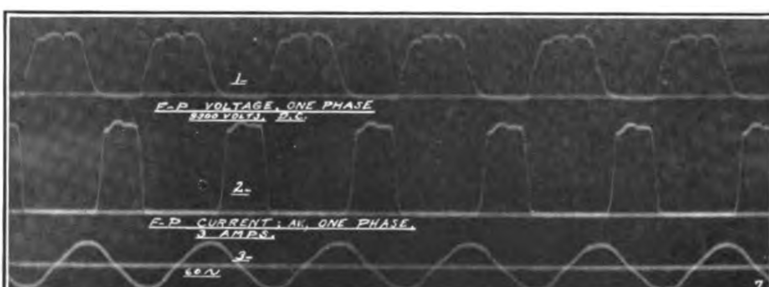


FIGURE 31—Oscillogram Showing Current and Voltage for the Rectifier Tubes in One of the Six Phases

The filter consists merely of a series inductance in the neutral and a large capacity in shunt with the load. The ripple frequencies in the output of a balanced system of ϕ phases are multiples of ϕf, where f is the frequency of the power supply. Their approximate amplitudes are readily found from the expression[9]

[9] A similar expression has been derived by D. C. Prince and probably by others who have worked with polyphase rectifiers. See this journal, October, 1922.

$$\text{ripple voltage} = \frac{2}{\phi^2 n^2 - 1} \times \text{average voltage.}$$

This expression refers to the voltage applied to the filter. It does not apply under some conditions of light load when the current in the neutral is interrupted during a portion of the cycle. For any assumed filter design the percentage of ripple applied to the load may thus be readily calculated. This procedure was employed in the design of the filter for the single-phase rectifier previously described and for the six-phase source which we are now considering. The method has been found to be of value. In the present case the 360-cycle ripple frequency is predominant. Its amplitude as applied to the filter is 5.7 percent. The filter employs an inductance of one-half henry and a capacity of three microfarads. With these constants the 360-cycle component is reduced to about 0.8 percent, while still higher frequencies are attenuated to values which are negligibly small for most purposes.

The general question as to how much ripple is permissible in the power supply of the last stage has considerable general interest, particularly since the requirements of the ordinary system and those for single-sideband transmission differ so radically. In the former the disturbance manifests itself primarily as a continuous note of the same frequency. Its presence is particularly objectionable during lulls in the modulation. In the latter system there is no effect in the absence of modulation, since there is then no radio frequency to act as a carrier. It is only during modulation that the disturbance appears, and then it is due less to the ripple tone itself than to components the frequencies of which are those of modulation plus and minus that of the ripple. It has been found experimentally at Rocky Point that with the six-phase rectifier the presence of the normal ripple in the plate supply was not a serious matter. Calculations indicate that even with a four-phase system this distortion would be permissible for many purposes, but that one of only three phases would not be satisfactory without a filter.

The second of the general requirements enumerated above relates to the impedance which the power source presents to the flow of the low frequency currents generated in the amplifier. It has been indicated that this consideration is of no importance in the case of a radio frequency amplifier of the first class, since for such there are no low frequency components in the anode current. When, however, we pass to class two and particularly to class three this question becomes one which must be considered, for unless precautions are taken, distortion may be introduced

and abnormal voltages may be produced in certain parts of the system. Since in the case of single-sideband telephony the power source ripple is not as troublesome as in the usual system, investigation may prove that it is best to reduce or omit certain elements of the filter if they are such as to increase the impedance to load circuit currents. In estimating this quantity in the case of a rectifier, we are at once confronted by the complication that it depends upon the load, on account of the curvature of the tube characteristics, and that it differs in different parts of the cycle of the supply current, because the number of phases carrying the load is not constant. However, the order of magnitude of the impedance may be ascertained by assuming simple conditions. For instance, in the six-phase rectifier discussed it might be postulated that one and only one of the phases of each delta carry current at a time. The tube impedance would be fixed by the fact that each of these phases would then carry half of the load. It is desirable that this impedance be small compared with the resistance of the load into which the rectifier delivers power at all frequencies that are apt to arise in the amplifier.

The above-mentioned impedance requirement is usually easier to meet in the system employing carrier and both sidebands than in the single-sideband system. In the discussion of amplifiers of the third class as applied to radio telephony, it was shown that in the case of the former the predominant frequencies generated in the amplifier are those of the voice signal itself, and this is strictly so if the anode direct current is proportional to the effective grid input. Hence, it is not far wrong to say that large currents the frequencies of which are lower than those of the signal are not present. There is thus a fairly definite lower limit below which the filter impedance may be high. On the other hand, with a single-sideband the low frequency anode currents are determined by differences of speech frequencies and therefore the impedance must be low for all from zero to a frequency numerically equal to the band width transmitted.

Of the three general requirements stated above, the last relates to regulation. When a third class amplifier is used in the ordinary radio telephone system, the fluctuations in the direct current load due to changes in modulation are usually not very large. The opposite is true when the carrier is eliminated, and therefore the decrease in voltage which results when the load is increased should be made small for obvious reasons. It is therefore important in designing a rectifier system for this purpose to use tubes of low resistance, to make the transformer leakage

reactance as small as permissible, and to employ means which allow the currents in the various phases to flow for a large fraction of a cycle. It is usually desirable in the design of the transformer to make the coupling between the whole of the primary and each half of the secondary as close as possible.

A photograph of the six-phase rectifier unit is shown in Figure 32. Twelve two-element tubes, two in parallel for each phase, are employed. Their anodes are mounted in jackets which are connected thru means for measuring the anode current and thru protective relays to earth and are hence substantially at earth potential. This practice has been followed in all the rectifiers employing water-cooled tubes on account of the simplification of design that results. The large individual hose coils which would otherwise be necessary are dispensed with. The jackets, meters and protective relays need not be mounted on high voltage insulators. It becomes necessary, of course, to provide separate filament transformers for each phase, and insulation between primary and secondary must be increased to withstand the higher voltages. However, the problem of obtaining suitable air-cooled transformers is not a serious one when the direct potential required is 15,000 volts or less. The filament transformers are mounted directly over the rectifier tubes. A thermometer for measuring the outlet temperature is mounted above the glass window in the front of the set. Below is a board with twelve jacks, one corresponding to each tube. When the plug is inserted the corresponding anode current is read on the small ammeter at the right. Under these jacks are the twelve relays which afford protection against an excessive load in any tube.

Above the rectifier set, Figures 14 or 32, can be seen six "pancake" choke coils. Their purpose is to prevent injury to the main transformer as a result of surges, such as might be set up by the failure of a rectifier tube. The six spark gaps are in series with resistances consisting of short pieces of hose containing a dilute solution of potassium nitrate. One terminal of each gap is connected to the filaments of one phase. The other terminal is connected to earth thru the resistance and a relay. The latter is common to the six gaps. These spark gaps and the relay are so adjusted as to operate when an excessive voltage surge is set up. The opening of the relay breaks the control circuit and shuts off the power. The objection may be raised that such an arrangement will result in frequent interruptions of service which might be avoided by the use of lightning arresters which relieve the surge and then open the circuit. Experience, however, has shown

that interruptions due to this are rare and that when they do occur they are usually accompanied by a disturbance elsewhere in the system which of itself would be sufficient to cause an interruption.

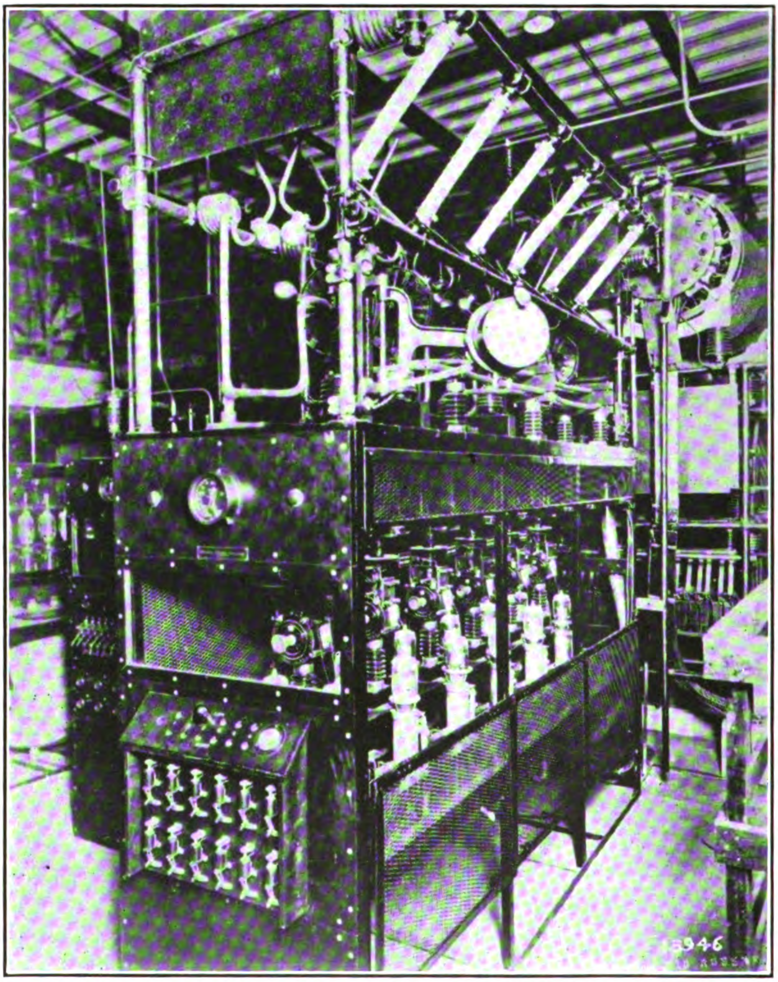


FIGURE 32—Six-phase Rectifier

Figure 33 shows the three-microfarad smoothing condenser, capable of withstanding 17,000 volts. It is made up of about 2,600 standard paper telephone condensers.

In Figure 34 is shown the 300 kilovolt-ampere 22,000-volt power transformer. The six terminals of the three-phase primary

are brought out separately to permit of either delta connection for normal, or star for reduced potential. There are seven bushings for the secondary, one for each phase and one for the neutral.

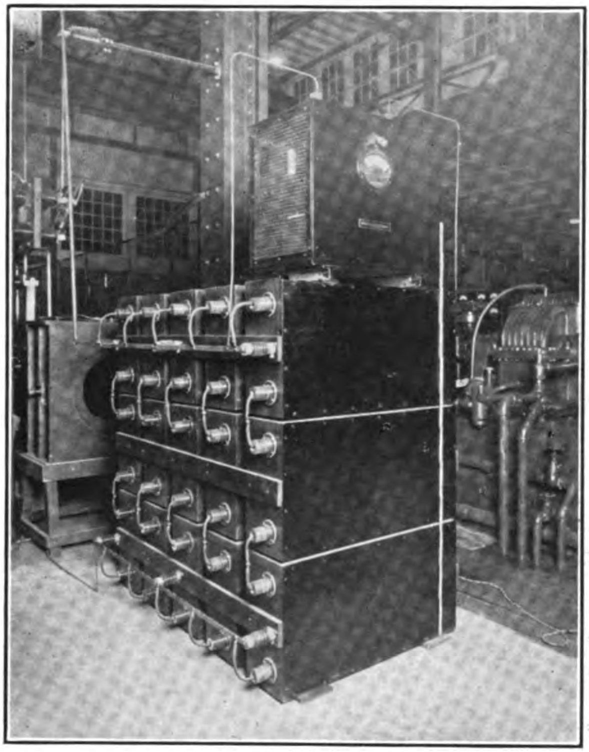


FIGURE 33—3-microfarad Smoothing Condenser

OVER-ALL SYSTEM

The complete amplifier system is shown in Figure 35. This includes the power apparatus and the control circuits. Figure 36 is an overall frequency-response characteristic for the entire power amplifier beginning with the input circuit of the first stage and ending with the total antenna current. The curve was taken for the condition of constant input voltage applied to the first stage at the maximum signal amplitude.

For purposes of comparison, the antenna resonance curve is shown dotted. It will be noted that the transmission band width of the antenna alone is about one-half that of the whole combination. This is a good example of the possibilities of band width improvement by properly proportioning circuits. However, it will be recalled from the discussion of the output circuit,

that such expansion of band width is accomplished at the expense of the vacuum tube capacity. Therefore, the kilovolt-ampere rating for the tubes must be somewhat greater than the maximum kilowatt output. Since the antenna systems for trans-

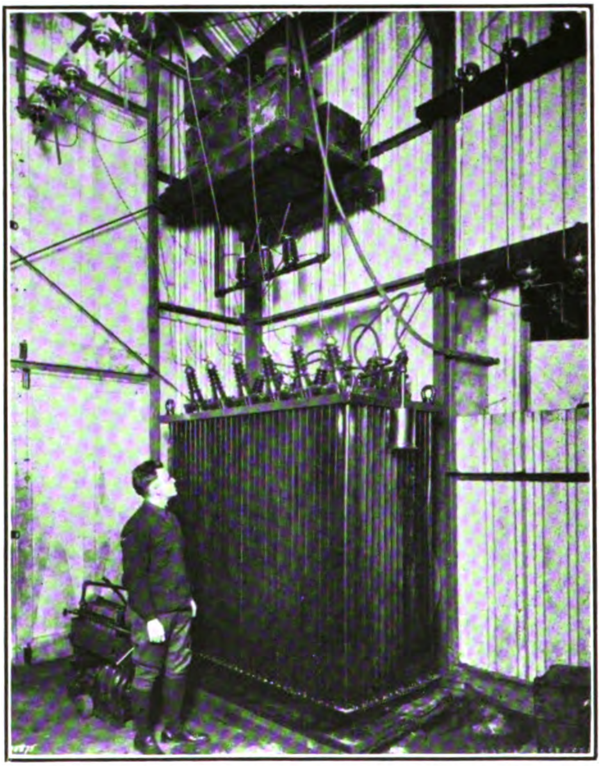


FIGURE 34—Six-phase 300-kva. 22,000-volt Transformer and Filter

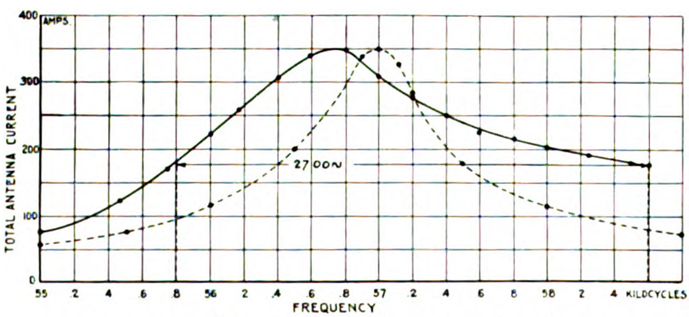


FIGURE 36—Over-all Transmission Characteristic. The resonance curve of the antenna is shown in dotted lines

oceanic telephone stations are expensive structures, it is obvious that for a specified transmission band width the most economical system will be one in which tube replacements and power costs are balanced against antenna carrying charges.

In closing, the writers wish to acknowledge the valuable engineering assistance of Messrs. H. R. Knettles, J. P. Schafer, M. E. Fultz, and E. J. Sterba, who participated in the design, construction, and experimental operation of the system. We are also indebted to Mr. C. W. Hansell, of the Radio Corporation of America, whose experience with high power radio plants made his many suggestions of great value to us during the construction period. We are also indebted to the operating personnel at Radio Central station for their excellent and whole-hearted co-operation thruout the work.

Research Laboratories of the American Telephone and Telegraph Company, and Western Electric Company, Incorporated, New York City.

June 7, 1924.

SUMMARY: The paper describes the development of a 150-kilowatt (output) radio frequency amplifier installation built for trans-atlantic telephone tests. The characteristics of the single-sideband eliminated-carrier method of transmission are discussed with particular reference to its bearing upon the design of the power apparatus. A classification of amplifiers is proposed in which there are three types distinguished from each other by the particular portion of the tube characteristic used. The water-cooled tubes employed in these tests are briefly described, special consideration being given to their use in a large installation. The system is then shown in outline by means of a block diagram, the elements of which are subsequently discussed in greater detail. The theory, electrical design, and mechanical construction of the last two stages of the amplifier are outlined, including the output and antenna circuits. Means employed to prevent spurious oscillations are described. The method used in increasing the transmission band width to a value much greater than that of the antenna is explained. The power requirements of a single sideband installation are outlined and a description of the six-phase rectifier used as a source of high potential direct current is given, together with a brief theoretical treatment of its operation. Circuit diagrams, photographs, and a number of characteristic curves are discussed.

RE-RADIATION FROM TUNED ANTENNA SYSTEMS*

BY

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INTRODUCTION

A tuned receiving antenna which is in the field due to a distant transmitting station, produces, by reaction, disturbances in the magnitude and direction of that field in the immediate vicinity of the receiving antenna. These reactions become important in studies of re-radiation and of the effects which they produce upon nearby coil antennas used for direction-finding purposes. The following investigation was undertaken to determine the magnitude and characteristics of such reactions for cases of reception wherein local oscillation does not occur. The results have led to a rather simple method of measurement of the equivalent height of an antenna system.

THEORETICAL CHARACTERISTICS

RADIATION FORMULA

It has been shown by J. H. Dellinger[1] that the vertical component of the field strength at the earth's surface at any point due to a current flowing in a flat-top antenna is:

$$H = \left[\frac{2\pi}{\lambda} \frac{h\,I}{10\,d} + j \frac{h\,I}{10\,d^2} \right] \alpha \tag{1}$$

where

H = field strength
h = antenna height
d = distance from antenna
λ = wave length
I = antenna current in vertical portion
α = absorption factor.
$j = \sqrt{-1}$

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*Received by the Editor, December 5, 1924.

[1] "Principles of Radio Transmission and Reception with Antenna and Coil Aerials," J. H. Dellinger, S. P. of Bureau of Standards, Number 354, 1919.

[illegible] per centimeter when [illegible] in centimeters. [illegible] in this paper.

[illegible] a function of the wave length, [illegible] the season of the year [illegible] for day-time transmission over sea water being

$$a = [illegible] \qquad (2)$$

The first term of equation (1) represents the radiation field, and the second term the induction field, due to the current in the vertical portion of the transmitting antenna. At small distances from the antenna the induction field predominates, and at large distances the radiation field predominates, the two fields being equal at the distance $\frac{\lambda}{2\pi}$.

The field, equation (1), will induce a voltage in a receiving antenna:

$$E = v h_2 H [illegible] \qquad (3)$$

where

E = induced voltage

v = velocity of propagation of field

h_2 = height of receiving antenna

The current which will flow in the receiving antenna at resonance is then:

$$I_r = \frac{E}{R} \qquad (4)$$

where

R = effective resistance of receiving antenna.

These expressions have been derived by Dellinger on the following assumptions:

(1) The current distribution in the vertical portions of both transmitting and receiving antennas is uniform.

(2) The flat top contributes nothing to the field at a distance from the antenna.

(3) The earth's surface is non-conducting, that is, no image of the antenna is formed.

(4) The distance from the antenna is large compared with the antenna height.

Under practical conditions, the current is not strictly uniform throut the vertical portion of the antenna, particularly in short wave antennas where the flat-top is small. The fact that

[4] "Quantitative Experiments in Long Distance Radio Telegraphy," L. W. Austin, "Bulletin Bureau of Standards," 7, October, 1911, page 315.

the earth is a partial conductor, and a partial image may therefore be formed, tends, as pointed out by Dellinger, to offset the first discrepancy. It is impossible, however, to take account mathematically of the effects of masts, trees and buildings upon the radiating properties of the antenna, and these factors introduce some uncertainty into calculations.

In practice, the fact that the current distribution is not uniform is taken into account by assigning an effective height to the antenna, this height being defined[3] as the product of the maximum height of the antenna above the ground, and the form factor. The form factor is defined[3] as the ratio of the average value of the current thruout the vertical portion of the antenna to the maximum value of the current in that portion. By using the effective height in the radiation formulas, however, we have still disregarded the effects of any nearby objects upon the field strength, and also the effects due to the fact the earth's surface may be a partial conductor and hence may form a partial image of the antenna.

It seems desirable, therefore, to define a term—the *equivalent height* which may be obtained by measurement, and which will take into account all of the above-mentioned factors. The *equivalent height* of an antenna will be defined as the height of an ideal antenna having a current equal to that at the base or anode of the actual antenna. The ideal antenna is considered as a vertical antenna with a uniform current distribution, erected on a non-conducting surface with no masts or other nearby structures. The equivalent height, which will be designated by h', is then that height, which, if substituted in the above formulas derived for the ideal antenna, will make the observed values of field strength agree with the calculated values. If this height is determined by observation near an antenna, the field strength at any point may be determined by calculation if the attenuation due to the absorption by the earth's surface is known. In an antenna system having directional properties, the observed equivalent height will be different in various directions, whereas the effective height as defined above does not determine this property. The observed equivalent height may be greater than equal to or less than the effective height, depending upon the magnitude and nature of the disturbances produced by the flat-top, the semi-conducting earth's surface, and by nearby absorbing or re-radiating structures. The difference between the equiva-

[3] The Institute of Radio Engineers—"Report of the Committee on Standardization for 1922," pages 10 and 12.

lent and effective heights in any particular case will be a measure of the effects produced by these factors.

These same considerations also apply to the receiving antenna, and it is necessary to use the equivalent height in the equations for the total induced voltage.

Inasmuch as this discussion concerns particularly the field strength in the immediate vicinity of the antenna, it is now proposed to find corrections for the above radiation formula, equation (1), when the distance from the antenna is not large compared with the antenna height.

Field Strength Near An Antenna

At short distances from the antenna, the induction field predominates, as previously mentioned, and so we may calculate the correction factor for the induction field and assume that it applies also to the radiation field without serious error.

Case I. Vertical Antenna, Uniform Current Distribution

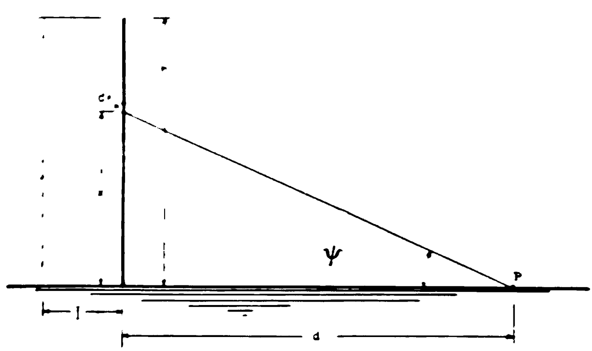


Figure 1—Vertical Antenna—Uniform Current Distribution

At the point P:

$$d\,H = \frac{I \cos \psi \, d\,x}{10(x^2+d^2)}$$

But
$$\cos \psi = \frac{d}{\sqrt{x^2+d^2}}$$

Hence
$$H = \frac{I\,d}{10}\int^{h} \frac{d\,x}{(x^2+d^2)^{3/2}}$$

or
$$H = \frac{I\,h}{10\,d^2} \cdot \frac{d}{\sqrt{h^2+d^2}}$$

the correction factor then being:

$$K_1 = \frac{d}{\sqrt{h^2+d^2}} \tag{5}$$

When d is large compared to h, K_1 becomes unity, and the expression for H reduces to that given for the induction field in equation (1). This correction is the equivalent to the form factor previously defined.

Case II. Vertical Antenna, Linear Current Distribution

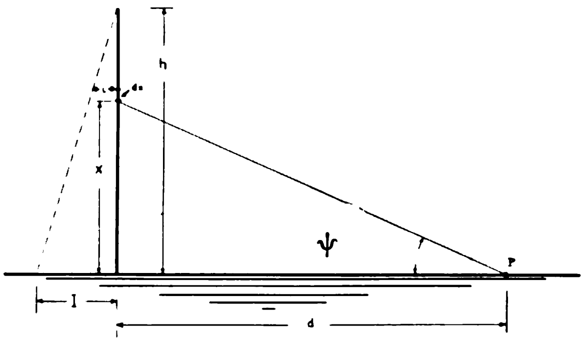


Figure 2—Vertical Antenna—Linear Current Distribution

At the point P:

$$d\,H = \frac{i \cos \psi \, d\,x}{10(x^2+d^2)}$$

But $$\cos \psi = \frac{d}{\sqrt{x^2+d^2}}$$

$$i = \frac{h-x}{h} I$$

Hence $$H = \frac{I\,h}{10\,d^2}\int_0^h \left[\frac{h}{(x^2+d^2)^{3/2}} - \frac{x}{(x^2+d^2)^{3/2}}\right] d\,x$$

or $$H = \frac{I\,h}{10\,d^2} \cdot \frac{d}{h}\left[\sqrt{h^2+d^2}-d\right]$$

The correction factor is then:

$$K_2 = \frac{d}{h}\left[\sqrt{h^2+d^2}-d\right] \tag{6}$$

which reduces to

$$K_2 = \frac{1}{2}$$

as d becomes large compared with h.

Case III. Flat-top Antenna Linear Current Distribution

height h' with uniform current distribution, the currents at the nodes being the same in both cases. Assume this antenna to be situated in the field from a transmitting antenna which is at such a distance that the field is substantially uniform in magnitude and direction in the immediate vicinity of the receiving antenna. Let H be the effective value of this field. Over a small area in the vicinity of the receiving antenna, H may be represented by a vector the direction of which is at right angles to the line pointing to the transmitting antenna (Figure 4).

This field will induce a current in the receiving antenna (equation (3)):

$$I_r = 3 \times 10^{10} \frac{h_r'}{R} H \times 10^{-8} \tag{10}$$

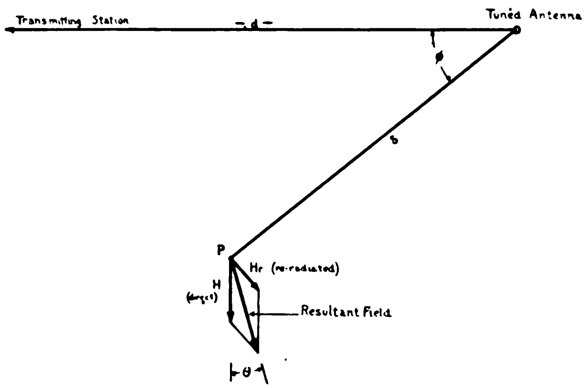


FIGURE 4—Field Distortion Due to Re-radiation

where h_r' is the equivalent height of the receiving antenna. Due to the current flowing in the vertical wires, the receiving antenna then becomes a miniature transmitting antenna, and a secondary or re-radiated field will be set up about it. The vertical component of this field at the earth's surface, and at a distance δ from the receiving antenna will be, by equation (1):

$$H_r = \left[\frac{2\pi}{\lambda} \frac{h_r' I_r}{10\, \delta} + j \frac{h_r' I_r}{10\, \delta^2} \right] K_1 \tag{11}$$

the induction field predominating at points near to the antenna. The absorption factor is of course unimportant for this case. The correction factor, K_1, used for proximity to the antenna is that for a vertical antenna with uniform current distribution, equation (5).

The resultant field strength at any point near to the receiving antenna will be a combination of the direct field, H, with the re-radiated field, H_r, with proper regard to the time and space phase relations of the two components. For points within the distance $\lambda/8$ of the receiving antenna, the two components will be closely in time phase with each other, and we may therefore make the simple space vector combination shown in Figure 4.

If this procedure is carried thru for a number of points in the immediate vicinity of the antenna, the resultant field is as shown in Figure 5, which has been calculated for the values shown. This distortion is similar to that produced in a uniform unvarying field by a wire carrying a direct current.[4]

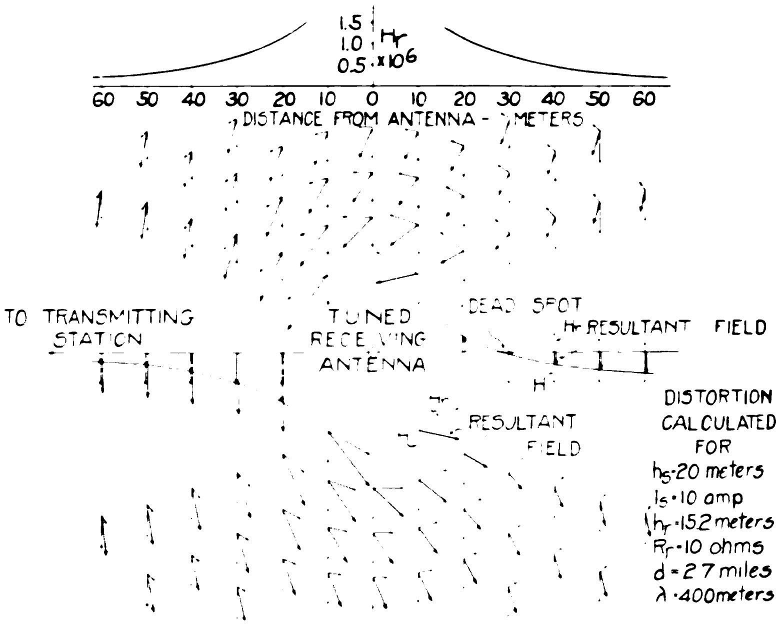


FIGURE 5—Calculated Field Distortion Near a Tuned Receiving Antenna

For points at greater distances from the receiving antenna, the phase relations due to time must be considered. The mathematical expression for the resultant field under such conditions will not be attempted here, but it may be pointed out that at a distance of $\lambda/8$ from the receiving antenna and between it and the transmitting antenna, the field component due to re-radiation

---

[4] See illustration, Figure XVII, in "A Treatise on Electricity and Magnetism," James Clerk Maxwell. Oxford, Clarendon Press, 1904. Volume II, 3rd Edition.

will be 90° out of phase with that due to the direct field. At right angles to the line of direction to the transmitting station, the components are 90° out of phase at a distance of $\lambda/4$ from the receiving antenna. On the side of the antenna opposite to the transmitting antenna, the two components are always in time phase with one another. The locus of the points of equal time-phase displacements for this case is a parabola, with its focus at the receiving antenna, and its major axis the line of direction to the transmitting antenna.

Characteristics of Field Near a Tuned Receiving Antenna

Observation of Figure 5, calculated for an ideal antenna of the constants indicated, and for distances within $\lambda/4$ from the receiving antenna, leads to several interesting facts. It is seen that in "front" of the antenna the field strength is considerably reinforced, due to the re-radiated component. To the "rear" of the antenna, and close by, the field is entirely due to the re-radiated component, and is therefore opposite in direction to the direct field. As the distance from the antenna in this direction increases, the field intensity passes thru a zero value, and becomes larger in the original direction, always, however, remaining less in magnitude than the original value of the direct field. At all other positions relative to the antenna, the resultant field is displaced at an angle to the direct field, and in general has a magnitude different from the original field.

Directly in the "rear" of a tuned receiving antenna we should then expect to find a position of zero field intensity, or "dead-spot" for the transmitting station to which the receiving antenna is tuned.

Comparison With Bureau of Standards Tests on Washington Monument

The Bureau of Standards have published[5] the results of tests made to determine the distortion produced by the Washington Monument in Washington, D. C., upon the field from a distant transmitting station. Maximum distortion was found to exist at a wave length of 800 meters, and it was assumed that this was the natural wave length of the monument considered as an antenna. The resultant field at 625 meters was also plotted and is herewith reproduced in Figure 6. It is very similar in general appearance to that calculated above, Figure 5. All distances

[5] "The Radio Direction Finder and Its Application to Navigation," F. A. Kolster and F. W. Dunmore, S. P. of Bureau of Standards, number 428, page 550, 1922.

shown in Figure 6 are within the distance $\lambda/4$ from the monument.

It should be noted that position number 6, directly to the "rear" of the monument is marked "dead." This is then the approximate position of the "dead-spot" noted above, and shown in Figure 5.

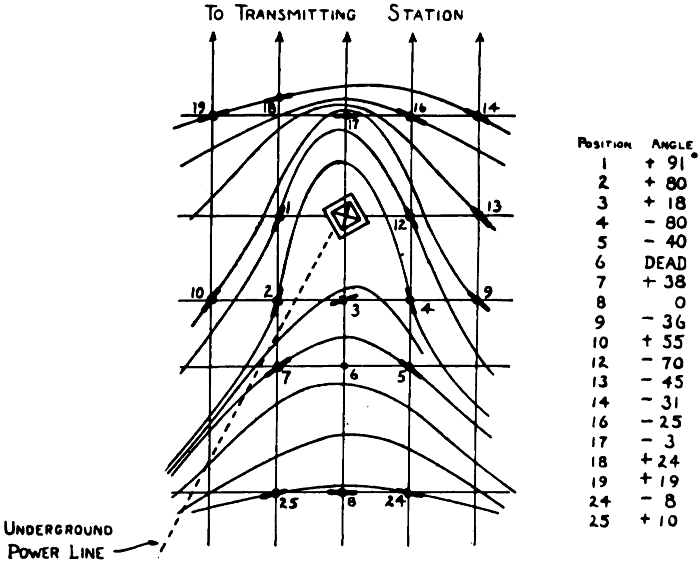


FIGURE 6—Field Distortion Caused by Washington Monument
(From S. P. Number 428: "The Radio Direction Finder and Its Application to Navigation," Bureau of Standards)

DISTORTION NEAR UNIVERSITY ANTENNA

A series of observations similar to those made by the Bureau of Standards were taken in the immediate vicinity of the antenna at the University of Minnesota. A small coil aerial of six turns on a form one meter square was mounted on a surveyor's transit, Figure 7, and observations were made on the apparent direction of a local broadcast station both with and without the receiving antenna being tuned. The broadcasting station (WLAG) is about two and one-half miles (4 km.) distant, and the field was therefore substantially uniform over the region investigated. The results of the measurements are shown in Table 1, which gives the observed directions together with the signal audibility before and after tuning the antenna to resonance. The angle of distortion and the audibility ratio are also given. Figure 8 shows the observed field distortion for the points taken, and the results are

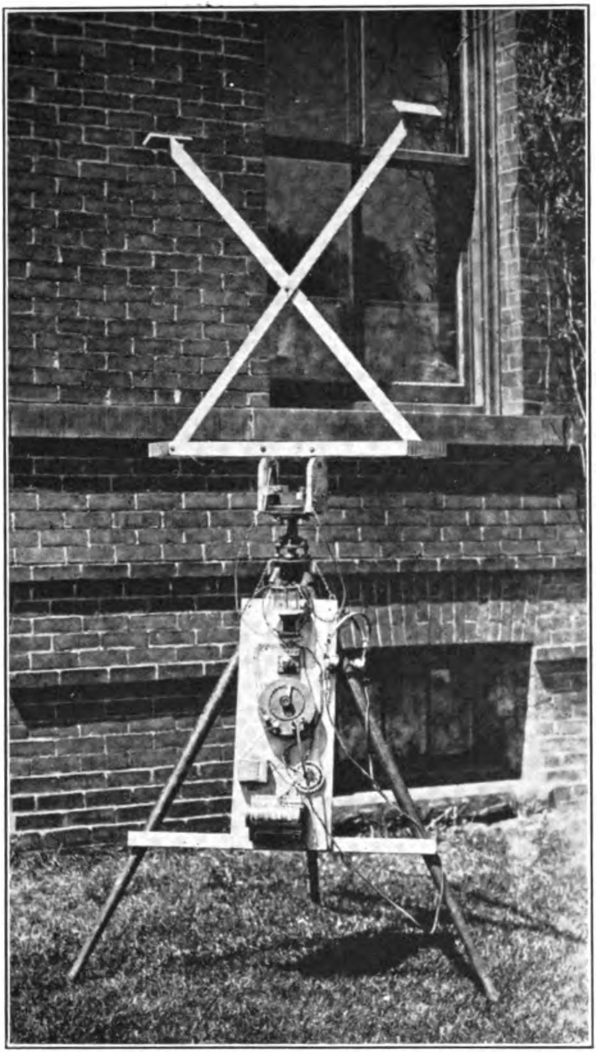


FIGURE 7

seen to be similar to those given in previous illustrations. The audibility measurements, while only very rough indications, show that the audibility is quite substantially increased in the direction of the broadcast station, and is reduced on the opposite side. The general position of the dead-spot is indicated. The presence of alternating current power wires prevented the taking of observations on the north side of the antenna.

TABLE I

FIELD DISTORTION PRODUCED BY TUNED ANTENNA AT UNIVERSITY OF MINNESOTA

Position (Figure 8)	Antenna Not Tuned				Antenna Tuned				Angle of Distortion	Audibility Ratio
	Minimum A	Minimum B	Apparent Bearing	Average Audibility	Minimum A	Minimum B	Apparent Bearing	Average Audibility		
A	N 74° E	N 110° W	N 108° W	15	N 61° E	N 116° W	N 117.5° W	18	9.5° S	1.2
B	N 82.5° E	N 97° W	N 97° W	30	N 30° E	N 147° W	N 148.5 W	350	51.5° S	11.7
C	N 85° E	N 96° W	N 95.5° W	350	N 83° E	N 101° W	N 99° W	800	3.5° S	2.3
D	N 64° E	N 111° W	N 111° W	50	N 69° E	N 107° W	N 109° W	80	2° N	1.6
E	N 78° E	N 110° W	N 106° W	40	N 52° E	N 123° W	N 125.5 W	4	19.5° S	0.1
F	N 50° E	N 130° W	N 130° W	12	N 50° E	N 124° W	N 127° W	10	3.0° N	0.8
G				0	No Minimum		N 95° W	4		
H	N 52° E	N 131° W	N 129.5° W	30	N 70° E	N 110° W	N 110° W	45	19.5° N	1.5
I	N 69° E	N 109° W	N 110° W	35	No Minimum		N 110° W	250	0	7.2
J	N 90° E	N 87° W	N 91.5° W	10				0	Dead Spot	0
K	N 70° E	N 111° W	N 110.5° W	15	N 60° E	N 118° W	N 119° W	35	8.5° S	2.3
L	N 75° E	N 103° W	N 104° W	6	N 63° E	N 115° W	N 116° W	12	12° S	2.0
M	N 72° E	N 105° W	N 106° W	7	N 52° E	N 124° W	N 126° W	15	19.5° S	2.1
N	N 60° E	N 118° W	N 119° W	80	N 57° E	N 122° W	N 122.5° W	120	3.5° S	1.5
S	N 78° E	N 99° W	N 100.5° W	22	N 87° E	N 90° W	N 91.5° W	40	9° N	1.8

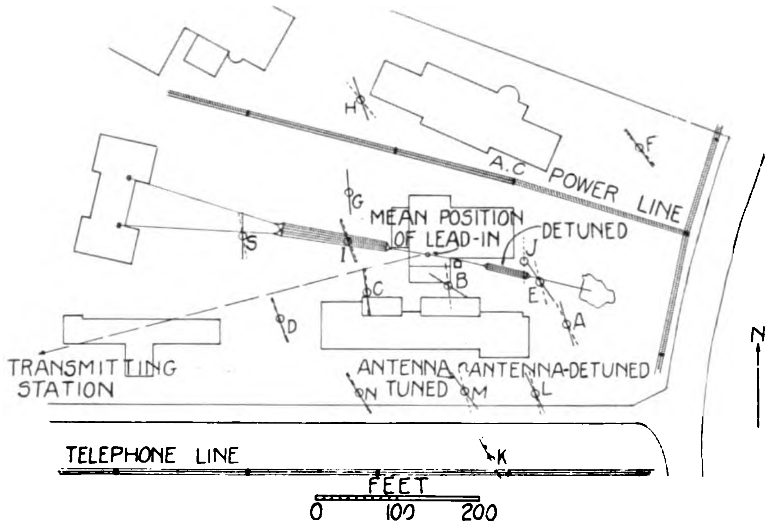


FIGURE 8—Field Distortion Due to Tuned Antenna at University of Minnesota

APPLICATION OF CHARACTERISTICS OF DISTORTED FIELD TO PRACTICAL USES

ELIMINATION OF INTERFERENCE

It is possible to utilize the "dead-spot" associated with a tuned receiving antenna for the purpose of eliminating interference from a powerful local station when it is desired to receive a distant station operating exactly on or near to the same wave length. The antenna is tuned to the interfering station and a coil antenna, or other antenna of small dimensions, is located in the resulting dead-spot, and used in the reception of the distant signal. All stations in the direction of the interfering station and operating on or near the same wave length are shielded from the coil antenna. Used in this manner, the antenna is merely a device for providing shielding for signals of one wave length and from one direction.

LOCATION OF COIL ANTENNAS FOR DIRECTION FINDING PURPOSES

The angle of deviation of the resultant field from the direct field has been shown to be the greatest in a direction approximately at right angles to the direction of the transmitting station. In locating coil antennas for direction finding purposes, this fact should be considered. If necessary to determine the direction of a station from a position near a large structure which is likely to produce distortion thru re-radiation, the direction-finder

should be placed directly between the structure and the transmitting station, as determined by repeated observation, relocating the position of the coil aerial after each observation.

BROADENING OF MINIMUM SIGNAL POSITION OF COIL ANTENNA

When using a coil antenna near a re-radiating structure, the re-radiated field component will not, in general, be in time phase with the direct field. In observing directions at a distance from the structure such that the re-radiated component is neither exactly in phase nor 180° out of phase with the direct field component, no zero position will be obtained because both field components will not pass thru the zero value at the same instant of time. The re-radiated field component will generally be small compared to the direct component, however, so that the only effect produced will be a broadening of the minimum signal position similar to that produced by the "antenna effect."[6]

TRANSMISSION WITHOUT LOCAL SOURCE OF POWER

The fact that the receiving antenna becomes a miniature transmitter when tuned to resonance with a distant transmitting station, led to a test to determine how far from the receiving antenna these re-radiated effects might be observed. It has been determined that a local broadcast station (WLAG) produced a total voltage of about 0.50 volts in the antenna at the University of Minnesota. When this antenna was tuned to resonance with this station, its total resistance was about eleven ohms, and the current was therefore about forty-five milli-amperes. An interrupter and a telegraph key were also included in the antenna circuit, and the interrupter was arranged to break the circuit at an audio frequency. When using a coil antenna and a crystal detector, the re-radiated signals from this antenna were clearly readable thru the modulation of the broadcast station up to a distance of about one-half mile (0.8 km.). When no modulation was taking place, it was possible to transmit readable signals a distance of three miles (4.8 km.) to a non-regenerative receiver connected to a small indoor antenna, in the direction indicated in Figure 9. This transmission used no power except that received from the broadcast station 2.7 miles (4.3 km.) distant. By inserting a carbon microphone directly in series with the tuning inductance in the receiving antenna, speech was successfully transmitted a distance of 1.3 miles (2 km.), as is also shown in

[6] "S. P. Bureau of Standards, number 354," J. H. Dellinger, 1919, page 487. "S. P. of Bureau of Standards, number 428," Kolster and Dunmore, 1922, page 541.

Figure 9. The signals from the interrupter were also clearly audible at this distance.

The possibilities of development of this system for short range work where simplicity of apparatus is essential are apparent.

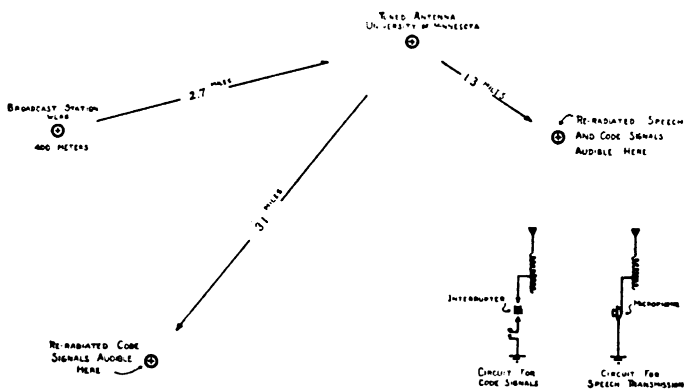


FIGURE 9—Transmission Without Local Power by Re-radiation

MEASUREMENT OF EQUIVALENT HEIGHT OF AN ANTENNA

Referring again to Figure 4 and recapitulating, consider a tuned antenna in the field of a distant transmitting station. Let the field intensity near the tuned antenna due to this transmitting antenna be designated by H. The current flowing in the vertical portion of the tuned antenna due to the voltage induced therein by the field H, will be, as above:

$$I_r = 3 \times 10^{10} \frac{h_r'}{R} H \times 10^{-8} \tag{10}$$

where h'_r is the equivalent height of the antenna. The tuned antenna then becomes a miniature transmitting antenna, producing a re-radiated field, the vertical component of which, on the earth's surface and at a distance δ from the antenna, may be expressed:

$$H_r = \left[\frac{2\pi h_r' I_r}{\lambda \; 10\,\delta} + j \frac{h_r' I_r}{10\,\delta^2} \right] K_1 \tag{11}$$

where

$$K_1 = \frac{\delta}{\sqrt{h_r'^2 + \delta^2}} \tag{5}$$

The magnitude of this re-radiated field is:

$$H_r = \frac{h_r' I_r}{10\sqrt{h_r'^2 + \delta^2}} \sqrt{\frac{4\pi^2}{\lambda^2} + \frac{1}{\delta^2}} \tag{12}$$

which, by substitution of equation (10) becomes:

$$H_r = \frac{30\,h_r'^2}{R\sqrt{h_r'^2+\delta^2}}\sqrt{\frac{4\pi^2}{\lambda^2}+\frac{1}{\delta^2}}\,H \tag{13}$$

Assuming that δ is sufficiently small so that the two components are practically in time phase with each other, we may write, by geometry from Figure 4, the following expression relating H and H_r:

$$H_r = H\frac{\sin\theta}{\sin(\phi-\theta)} \tag{14}$$

Substituting (14) in (13), we have:

$$\frac{\sin\theta}{\sin(\phi-\theta)} = \frac{30\,h_r'^2}{R\sqrt{h_r'^2+\delta^2}}\sqrt{\frac{4\pi^2}{\lambda^2}+\frac{1}{\delta^2}} \tag{15}$$

which is an expression relating h_r' with the distance δ, the wave length λ, the total antenna resistance R, and the angles ϕ and θ. A rather simple method of determing the equivalent height of any antenna thus arises.

The above expression, equation (15), cannot be solved explicitly for h_r' in simple form, but since h_r' enters the factor $\sqrt{h_r'^2+\delta^2}$ as a correction, its value may be estimated and a preliminary solution made. A second calculation using the new value is generally sufficient to determine h_r' within the accuracy of the method.

Solving, then:

$$h_r' = 0.1826\sqrt{\frac{R\sin\theta}{\sin(\phi-\theta)}}\sqrt{\frac{h_r'^2+\delta^2}{\frac{4\pi^2}{\lambda^2}+\frac{1}{\delta^2}}} \tag{16}$$

the value of h_r' on the right-hand side to be estimated as a first approximation. Any one set of units may be used for all lengths.

Equation (16) permits the measurement of the equivalent height of any antenna, by noting the apparent change in direction of a distant transmitting station as observed by a coil aerial near the antenna to be measured, as the antenna to be measured is tuned to resonance with the distant transmitting station. The other factors required for the determination are the distance of the coil aerial from the antenna to be measured, the wave length, the total antenna resistance, and the angles ϕ and θ of Figure 4.

In order that an angle θ may exist, it is necessary that the angle ϕ be made different from 0° or 180°. In order that θ may be large, the angle ϕ should be near 90°, and preferably somewhat larger than this value. If ϕ is made 90° for the determination, the factor $\frac{\sin\theta}{\sin(\phi-\theta)}$ of equation (16) becomes $\tan\theta$, and we have,

if $\phi = 90°$:

$$h_r' = 0.1826\sqrt{R \tan\theta}\sqrt{\frac{h_r'^2 + \delta^2}{\frac{4\pi^2}{\lambda^2} + \frac{1}{\delta^2}}} \qquad (17)$$

Any transmitting station may be used as a source for the measurement, and the results are independent of the characteristics and distance of the transmitting station, provided the distance is sufficiently great so that the field is uniform near the antenna to be measured. The coil antenna may be located up to a distance of somewhat less than $\lambda/4$ from the vertical portion of the antenna to be measured, at approximately right angles to the direction of the transmitting station relative to the antenna.

If necessary to determine the equivalent height for various directions from the antenna, the use of more than one transmitting station may be necessary. The equivalent height thus determined is that for the particular wave length used.

It should be noted that this measurement does not take into account the effect of the current in the antenna flat-top upon the field strength at the point in question. Any such effect, altho usually small, will be credited to the lead-in by using this method of measurement, and will cause an apparent increase or decrease in the apparent equivalent height according as this effect adds to or subtracts from the field component due to the vertical portion. If observations are made at increasing distances from the antenna, the observed values of equivalent height will either increase or decrease slightly with distance, and asymptotically approach the true value due to the current in the vertical portion alone, since at great distances there will be no vertical field component due to the current in a horizontal wire.

This method is more particularly applicable to measurements of the equivalent height of antennas of symmetrical and definite characteristics, such as the vertical types, the inverted L and the T. Antennas which do not have a strictly defined vertical portion are difficult to measure because of the uncertainty of the determination of the distance of the coil antenna from the vertical portion.

In order to secure accuracy in the measurement, the angle θ should be of the order of 10°. Smaller values are difficult to measure accurately, and with larger values the value of the function $\frac{\sin\theta}{\sin(\phi - \theta)}$ is changing rapidly. The distance δ should be as large as possible consistent with remaining within the region wherein the direct and re-radiated field components are sub-

stantially in time phase. This means that the angle ϕ should preferably be somewhat greater than 90°, say in the neighborhood of 110°. The total antenna resistance, R, must include the resistance of the tuning devise.

While not a particularly admirable method of measurement, largely because of the necessity for making the observations so close to the antenna, it is quite practical for approximate results. It may be noted that the accuracy of the calculation is quite good, due to the fact that the equivalent height enters into the equation twice, once as a receiving factor, and again as a re-transmitting factor. Any errors in the measurement of R, θ, ϕ or δ then show only approximately as the square root in the resulting values of equivalent height.

Application of Method

This method of measurement of equivalent height has been applied to the data of Table 1 and Figure 8, which was obtained near to the antenna at the University of Minnesota. From the data for position M, which is closely at right angles to the antenna relative to the direction of the transmitting station, and is at a distance of 170 feet (51.8 meters) from the lead-in, we find by equation (17):

$$h_e'{}_M = 16.7 \text{ meters} = 54.8 \text{ feet.}$$

From the data of position K, where $\delta = 263$ feet (80.2 meters) and $\phi = 90°$,

$$h_e'{}_K = 15.0 \text{ meters} = 49.2 \text{ feet.}$$

Similarly at position L, where $\delta = 217$ feet (66.2 meters) and $\phi = 114°$, we have by equation (16):

$$h_e'{}_L = 15.6 \text{ meters} = 51.2 \text{ feet.}$$

The true equivalent height for this wave length and for this direction is probably about 49 feet (15.0 meters). Position M was entirely cut off from the antenna by the building shown in Figure 8, and at position K only the top of the antenna was visible.

Current measurements in a coil aerial near to this antenna, and based on the radiation formulas of the early part of this paper, give the following values for the equivalent height:

South — 16.7 meters (= 54.8 feet)

Southwest — 16.0 meters (= 52.5 feet)

These figures are for a wave length of 279 meters, whereas those given above are for the 400-meter wave length.

It is interesting to note that the application of this equation

to the data given in Figure 6 of the Bureau of Standards' observations near Washington Monument, give the following values for the equivalent height of that structure:

Position number 9

$$h_r' = 13.5\sqrt{R} \text{ meters}$$
$$= 44.3\sqrt{R} \text{ feet}$$

Position number 13

$$h_r' = 16.2\sqrt{R} \text{ meters}$$
$$= 53.0\sqrt{R} \text{ feet}$$

Position number 14

$$h_r' = 14.2\sqrt{R} \text{ meters}$$
$$= 46.5\sqrt{R} \text{ feet}$$

Pierce's Tables[7] give the radiation resistance of a vertical antenna at the fundamental wave length as 36.6 ohms. Considering that the wave length of 625 meters is about 20 percent off resonance, a total impedance of the order of 50 ohms seems probable for this structure. If this value be assumed, the numerical values of the equivalent height determined above become:

Position number 9

$$h_r' = 313 \text{ feet}$$

Position number 13

$$h_r' = 374 \text{ feet}$$

Position number 14

$$h_r' = 328 \text{ feet}$$

$$\text{Average} = 338 \text{ feet}$$

The other observations cannot be used because of the disturbance produced by the underground cable. Since the total height is 532 feet, the form factor is 0.635 by this determination, which is rather too close to the theoretical value for a vertical antenna with sinusoidal current distribution, $2/\pi$ (0.637), to be entirely convincing. At any rate, the equivalent height of this structure is apparently closely equal to the effective height, a fact that would be expected, due to the absence of a flat-top and any nearby structure.

ACKNOWLEDGMENT

The writer wishes to express his appreciation of the assistance and co-operation of Professor C. M. Jansky, Jr., Mr. R. A. Braden, and others of the faculty of the Department of Electrical Engineering of the University of Minnesota, in the conduct of these experiments.

[7] "Electric Oscillations and Electric Waves," G. W. Pierce, McGraw-Hill, 1920, page 509.

SUMMARY: This paper considers Dellinger's treatment of the radiation from antenna systems, and applies correction factors thereto for the field strength near to the antenna. The term "equivalent height" is suggested and is differentiated from effective height. The expressions are then applied to the case of a tuned receiving antenna to determine the distortion produced in the field of a distant transmitter by re-radiation from the receiving antenna. The calculated results are then compared with the field distortion found by the Bureau of Standards near the Washington Monument, and also with measurements made near a typical antenna. The distorted field is found to have practical application as a means of eliminating interference; in short distance transmission without a local source of power; and in providing a means of measurement of the equivalent height of an antenna system.

NOMENCLATURE

The following nomenclature has been used in this paper:

H = Field intensity due to a transmitting antenna.

H_r = Field intensity due to a re-radiating antenna.

h = Maximum height of an antenna.

h_r = Maximum height of a receiving antenna.

h' = Equivalent height of an antenna.

h_r' = Equivalent height of a receiving antenna.

x = Height to a differential element in vertical portion of an antenna.

ψ = Vertical angle of differential element above ground as observed at a distance d.

I = Maximum current in vertical portion of antenna.

I_r = Maximum current in vertical portion of a receiving antenna.

i = Current at height x in vertical portion of an antenna.

d = Distance from vertical portion of an antenna.

δ = Distance from vertical portion of a re-radiating antenna.

E = Total induced voltage in a receiving antenna.

R = Total effective resistance of a receiving antenna.

b = Length of antenna flat-top.

K_1 = Distance correction factor, vertical antenna, uniform current distribution.

K_2 = Distance correction factor, vertical antenna, linear current distribution.

K_3 = Distance correction factor, flat-top antenna, linear current distribution.

c = Velocity of propagation of field from antenna.

θ = Angle of distortion of field due to re-radiation.

ϕ = Direction angle of observer relative to distant transmitting antenna.

α = Absorption factor.

ϵ = Constant, 2.718. . .

$j = \sqrt{-1}$

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

ISSUED MARCH 3, 1925—APRIL 28, 1925

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,520,129—E. H. Loftin and H. H. Lyon, filed April 5, 1920, issued March 17, 1925.

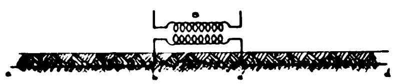


Fig. 2

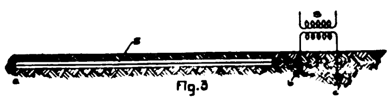


Fig. 3

NUMBER 1,520,129—Radio Signaling System

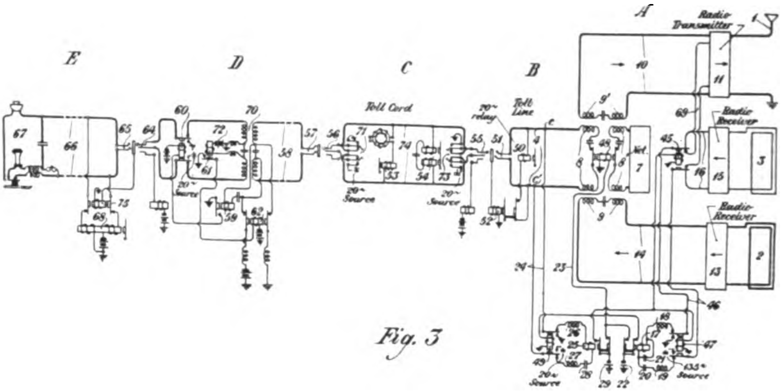
RADIO SIGNALING SYSTEM utilizing low extended antennas. The optimum length of the low horizontal antenna is described as being of the order of one-tenth the wave length of an operating signal and adjusted in resonance with the signaling wave. The invention is described with reference to underground and underwater antenna systems.

1,520,452—C. E. Willey, filed August 28, 1922, issued March 17, 1925.

RADIO DETECTOR of the cat-whisker crystal variety wherein a crystal holder is provided consisting of a cylindrical container adapted to receive a plug of soft metal in which a crystal is imbedded. The cylindrical container is open at both ends and fits over the head of a correspondingly shaped nut. In this manner, the crystal container is frictionally secured in position.

*Received by the Editor, May 12, 1925.

1,528,010—C. S. Demarest and M. L. Almquist, filed December 31, 1923, issued March 3, 1925. Assigned to American Telephone and Telegraph Company, New York.



NUMBER 1,528,010—Radio Signaling System

RADIO SIGNALING SYSTEM combining the advantages of line wire radio communication. Separate receiving circuits are provided for the reception of speech-modulated and signal-modulated carrier currents. The arrangement of the receiving circuits is such that a large amplification of the particular detected frequency or range of frequencies, which it is intended to secure may be received. The patent describes a terminal circuit which may be connected to a transmitting or receiving antenna with either speech currents or telegraph signaling currents. The switching means is actuated by signaling currents transmitted over the line for connecting the modulating circuit in desired relation to the antenna system and control circuits.

1,536,039—J. T. Bradley, filed April 3, 1923, issued April 28, 1925.

"NO-CAPACITY CONTROL" FOR CONDENSERS, in which the condenser plates may be moved for coarse adjustment and then moved for finer adjustment by means of gearing contained within the control knob secured to the condenser shaft.

1,528,011—C. S. Demarest and M. L. Almquist, filed December 31, 1923, issued March 3, 1925. Assigned to American Telephone and Telegraph Company, New York.

RADIO SIGNALING SYSTEM for slectivity signaling a particular station in a group of radio stations. In order that each station

of the system may be able to maintain telephone or telegraph communication with every other station and a particular station to the exclusion of other stations, a band of frequencies is employed in the system. Each receiving station includes a detector which is so tuned that it will almost oscillate when the frequencies of one of the bands are applied to the detector. The receiving circuits are thereby operated selectively, enabling the several stations in the system to communicate without interference.

1,528,032—S. A. Staege, filed January 14, 1921, issued March 3, 1925. Assigned to Westinghouse Electric and Manufacturing Company, Pennsylvania.

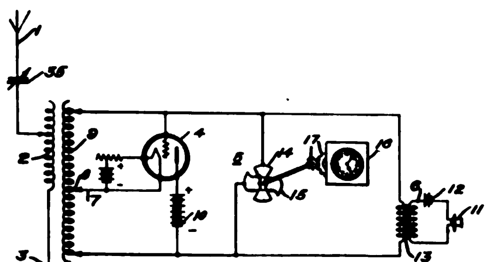


Fig. 1.

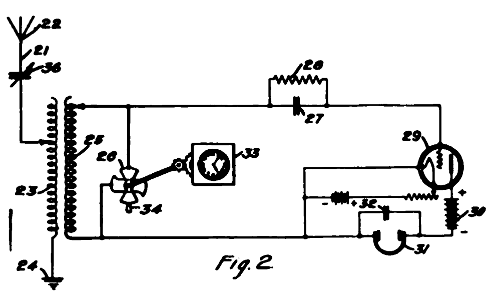
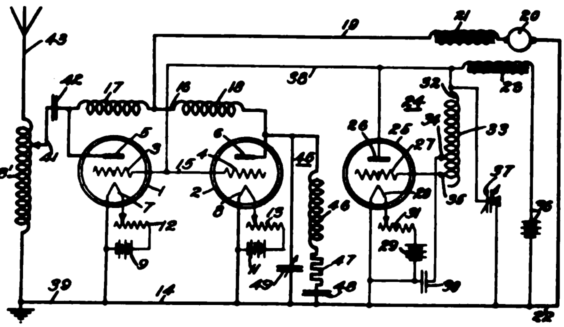


Fig. 2

NUMBER 1,528,032—Selective Signaling System

SELECTIVE SIGNALING SYSTEM, in which the carrier wave frequency of a transmitting station is caused to vary thru a predetermined cycle within certain limits, while the tuning of the receiving station passes thru the same cycle of variation in frequency. This is accomplished by means of a clock mechanism at both the transmitting and receiving station by which a variable condenser is cyclically moved simultaneously. The frequency at both the transmitting and receiving station may in this manner be uniformly controlled.

1,528,047—Frank Conrad, filed March 15, 1922, issued March 3, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,528,047—Radio Telephone System

RADIO TELEPHONE SYSTEM having a modulating system particularly adapted for high power operation. The modulation system employs a thermionic tube which is so associated with energy-absorbing circuits that only relatively small amounts of energy are dissipated in the modulator tube itself, the larger portion thereof being dissipated in the absorbing circuit. A pair of parallel connected valves are arranged in the transmitting circuit and modifications of the frequency of the impressed radio frequency currents cause opposite variations in the power absorbed by the respective circuits.

1,528,054—I. B. Harris, Wm. L. Harris, and E. B. Harris, filed March 24, 1922, issued March 3, 1925.

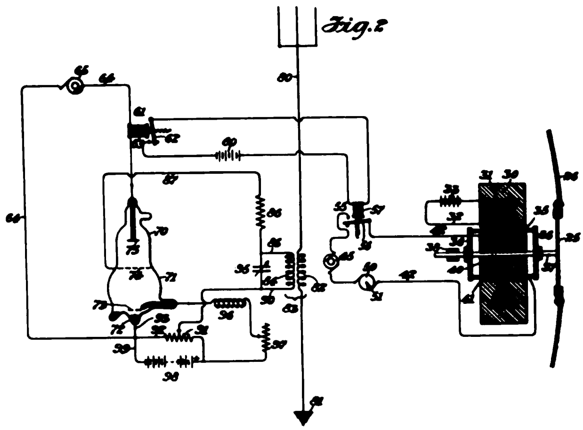
RECTIFIER tube having a plurality of independent cathodes anyone of which may be used in the event that another of the cathodes is destroyed. In this way the life of the rectifier bulb is increased, inasmuch as when one filament is destroyed another may be used. A form of connection for the separate cathodes is used in which a contact tap may be removed from the base of the tube to expose an auxiliary contact connected with another of the cathodes.

1,528,735—F. S. McCullough, filed June 2, 1924, issued March 3, 1925.

THERMIONIC-TUBE CONSTRUCTION, in which a radio frequency transformer is directly mounted within the tube adjacent the

tube electrodes. The transformer has a pair of inductively-coupled coils which have their terminals connected thru extremely short leads with the electrodes of the tube.

1,529,065—J. H. Hammond, Jr., filed December 11, 1916, issued March 10, 1925.



Number 1,529,065—System of Radio Control

System of Radio Control of submarine vessels and other movable bodies. A combined electromagnetic wave reception system and sound wave transmission system is illustrated, whereby received radio signals are caused to actuate a submarine compression wave sound transmission system for transmitting energy under water for control of sound receiving devices on board the submarine vessel.

1,529,096—H. C. Tucker, filed October 5, 1922, issued March 10, 1925.

Electrical Control Instrument, wherein a rotatable shaft is journaled with respect to a panel by means of a sleeve which is directly supported by the panel and forms a hub thru which the rotatable shaft may be passed. The invention is shown in connection with a variable electrical condenser, altho the claims set forth in the invention are generally applicable to various rotary instruments.

1,529,455—F. K. Vreeland, filed May 3, 1920, issued March 10, 1925.

Electrical Oscillator for securing purity of wave form in an oscillating circuit having inductance and capacity reactances

which are large with respect to the other impedances of the system with which it is associated. The specific means for securing this result includes a step-up coupling of the system including the power circuit to the oscillating circuit.

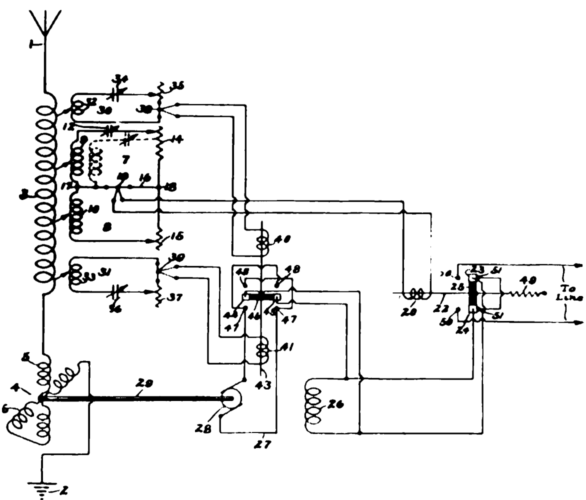
1,529,597—I. Langmuir, filed August 11, 1921, issued March 10' 1925. Assigned to General Electric Company, New York.

ELECTRON-EMITTING DEVICE AND METHOD OF PREPARATION of electron-emitting cathodes of the thoriated type, containing an amount of carbon sufficient to render the cathode less sensitive to the deleterious effect of gases than a cathode body unprovided with carbon. The specification describes a cathode which contains less than 3 percent by weight of carbon.

1,529,626—E. Y. Robinson, filed March 21, 1924, issued March 10, 1925.

VACUUM ELECTRIC TUBE for high powered operation, where the plate is sealed to the glass envelope and closes the lower portion of the envelope for the circulation of a cooling fluid around the depending metallic portion. The patent describes a method of removing gases and vapors from the metallic portion of the tube which consists in heating the metallic portion while protecting the zone at which the metallic portion is sealed to the glass envelope.

1,530,169—W. F. Grimes, filed June 7, 1923, issued March 10. 1925.



NUMBER 1,530,169 Radio Signaling System

Radio Signaling System, in which the signaling frequency of a transmitting station may be maintained constant. The antenna circuit of the transmitter is coupled with a pair of balanced circuits which at normal frequency remain in balanced condition. In the event that the frequency of the transmitter shifts, the control circuits become operative to adjust the antenna circuit to the normal frequency.

1,530,498—B. W. Kendall, filed November 20, 1917, issued March 24, 1925. Assigned to Western Electric Company, Incorporated, New York.

Synthesis of Compound Tones by Vacuum Tube Oscillators, by which various tones over the musical scale may be produced by electron tube circuits. A complex wave is generated which includes a plurality of different frequencies components which may be selected to produce desired oscillations of particular frequencies. The different frequencies may be utilized for the reproduction of audio notes of desired pitch and timbre.

1,530,660—W. F. Friedman, filed July 26, 1922, issued March 24, 1925.

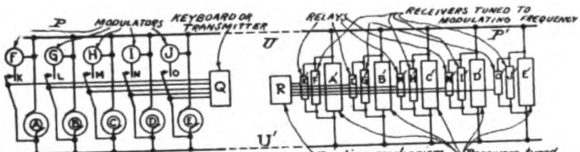


Fig.1

Number 1,530,660—Printing Telegraph System

Printing Telegraph System for effecting simultaneous transmission and reception of a plurality of code signal impulses representing the individual elements of the message characters that are to be transmitted. The object of the invention is to reduce the time necessary for the transmission of printed messages by telegraph. A set of generators of radio frequency oscillations at different radio frequencies are provided. A single modulation frequency is employed for telegraphically modulating the oscillations. A set of make and break keys are arranged to control the modulating or non-modulating of the oscillations. The keys are operated simultaneously and permutatively to correspond to the permutations of a plural unit signaling code representing message characters. At the receiving station sets of receiving instruments are arranged to isolate the respective oscillations, detect

the same, and control relay circuits which are operated permutatively to actuate a printing mechanism reproducing the transmitted messages.

1,530,[illegible]—J. H. Hammond, Jr., filed October 31, 1917, issued March 24, 1925.

MULTIPLEX SYSTEM FOR THE TRANSMISSION OF RADIANT ENERGY over the same antenna system. The transmitting circuit is supplied with energy from a single source of oscillations which is modulated by independent keying circuits, each of which may be interrupted at a rate within audibility. In this manner different tone frequencies may be produced for different signaling channels over the same antenna system.

1,53[illegible]84—J. O. Mauborgne and Guy Hill, filed June 29, 1921, issued March 24, 1925.

ANTENNA SYSTEM consisting of a flat spiral resonance wave coil. The wave coil is constructed of a fixed distributing inductance and capacity of such values that the spiral is capable of having developed thereon a number of standing waves corresponding to a wide range of frequencies. A point may be selected on the coil for connection of the wave so that the particular frequency [illegible] and connections taken to a radio transmitter [illegible]

[illegible] M[illegible], filed August 7, [illegible] issued March 24, [illegible]

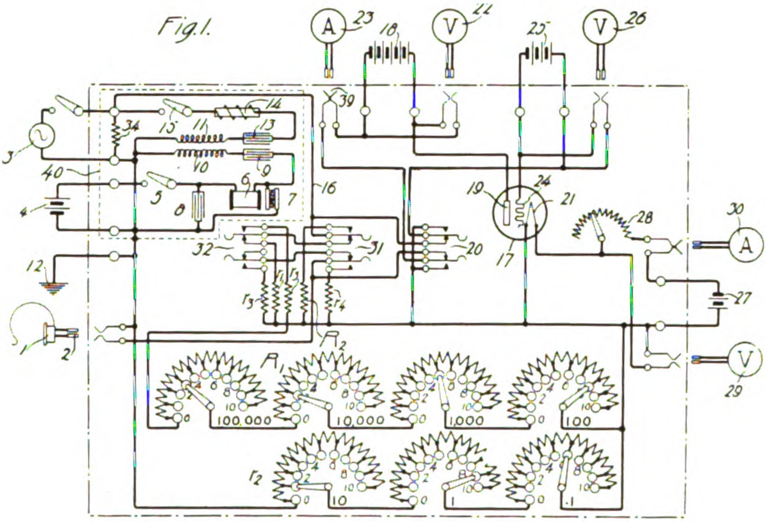
[illegible] contacts [illegible] filament [illegible] may be [illegible]

[illegible] March [illegible] Inter[illegible]

[illegible]

By

the present invention the constants of the tube can be read directly from dials which operate the different parts of the testing apparatus. An adjustable balancing resistance is provided connected in circuit with the electron tube tested. The resistance is calibrated in terms of the impedance of the space discharge path between two of the electrodes of the tube. When proper readings are obtained on meters in the tube circuits the tube constants may be determined from the dial settings.



NUMBER 1,530,988—TESTING VACUUM TUBES

1,531,029—F. M. Ryan, filed August 24, 1921, issued March 24, 1925. Assigned to Western Electric Company.

RADIO TRANSMISSION SYSTEM arranged to co-operate with a carrier-wave receiving system which has a closed energy receiving circuit or loop connected to the receiving apparatus with a circuit interposed between the receiving apparatus and the loop which has inductive reactance at desired frequencies functioning to exclude energy of undesired frequency from the receiving circuit. In this manner the transmission system may be operated in close proximity to the receiving system without interference therebetween.

1,531,278—G. H. Clark, filed February 16, 1921, issued March 31, 1925. Assigned to Radio Corporation of America.

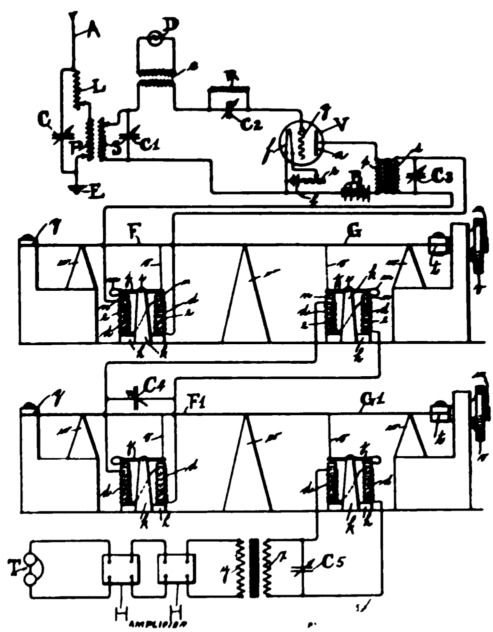
CONTROLLER for OSCILLATION GENERATORS of the arc type in which the striking of the arc may be controlled by an operator

at the transmitting station. The hydrogen gas supply to the arc is mechanically controlled before the arc is struck. A mechanical and electrical circuit arrangement is shown whereby the liquid which supplies the hydrogen gas in the arc chamber is first supplied to the chamber before the arc is struck.

1,531,633—H. J. Vennes, filed December 22, 1920, issued March 31, 1925. Assigned to Western Electric Company.

OSCILLATION GENERATOR having a circuit for production of currents of more than one frequency which are independent of one another. A plurality of tuned circuits are provided which are coupled with the oscillating current path. The tuned circuits may have the periodicity of any one of them adjusted, while the periodicities of any of the others of the tuned circuits may be maintained constant.

1,531,801—D. G. McCaa, filed May 2, 1922, issued March 31, 1925.



NUMBER 1,531,801—Signaling System

SIGNALING SYSTEM designed to eliminate the effects of static, strays, and electrical atmospheric disturbances. The invention is based on the principle that the telegraph signaling currents

have a characteristic note of tone whereas stray currents differ substantially from the audio frequency note of the signal. A mechanical vibratory system is provided at the receiving station which may be set into operation by the periodic signal resisting current. The vibration of the mechanical system will independently generate the signaling energy which is finally observed free and clear of the effects of static disturbances.

1,531,805—R. C. Mathes, filed July 10, 1920, issued March 31, 1925. Assigned to Western Electric Company.

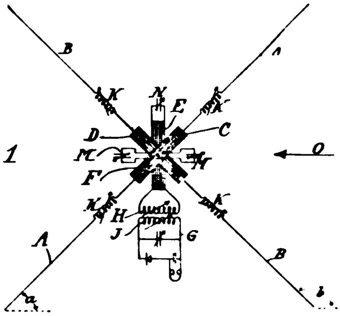
OSCILLATION GENERATOR having a regenerative feedback circuit which is not subject to the reaction of impedance of the output circuit. The object of the circuit arrangement is to produce a purer final wave form or more constant frequency than is produced in the usual electron tube oscillatory circuit. The tube of the present invention has a plurality of space discharge paths of different impedances and energy is supplied thru one path to control the regenerative operation of the circuit while load current is supplied thru the other path.

1,532,336—W. H. Nottage and T. D. Parkin. Filed December 18, 1920, issued April 7, 1925. Assigned to Radio Corporation of America.

RADIO TELEGRAPH CALLING DEVICE, in which a balance wheel is provided at a receiving station which has a natural oscillation equal to the interval between successive transmitted impulses so that tho the relay may be unaffected by ordinary signals, yet when a train of impulses at predetermined intervals is incident upon the receiver the oscillations of the relay are increased by the impulses sufficiently to enable it to actuate a bell or other signal or operate an electric corcuit for the control of apparatus.

1,532,356—R. A. Weagant, filed February 7, 1919, issued April 7, 1925. Assigned to Radio Corporation of America.

RADIO SIGNALING SYSTEM for minimizing the effects of static interference in radio reception. An antenna system is provided by which signaling energy and static energy may be received simultaneously from different directions with respect to the horizontal plane. The currents due to the static energy are balanced out in the receiving circuit while the currents due to the signaling energy are retained and amplified.

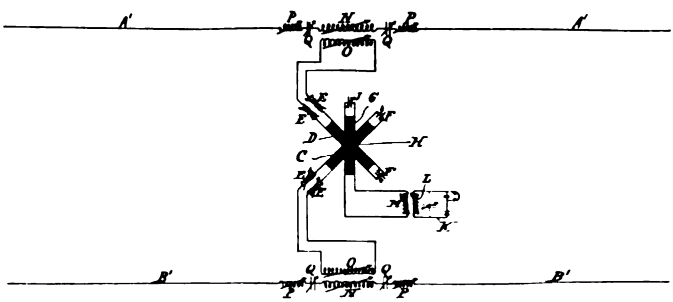


NUMBER 1,532,356—Radio Signaling System

1,532,364—O. A. Berman, filed October 2, 1922, issued April 7, 1925.

MANUFACTURE OF CONDENSERS, in which the condenser elements are constructed of conductive plates which may be moved positively towards or from an adjacent dielectric plate in an axial direction. A movable plate is mounted upon a screw-threading member which is adjusted in screw threads which pass thru an adjacent dielectric plate which separates the movable plate from the conductive plate.

1,532,367—R. A. Weagant, filed February 7, 1919, issued April 7, 1925. Assigned to Radio Corporation of America.



NUMBER 1,532,367—Method and Apparatus for Radio Signaling

METHOD AND APPARATUS FOR RADIO SIGNALING for reducing the effect of static disturbances. An antenna system comprising pairs of collectors is provided in which static energy may be successively received while signaling energy is simultaneously received from substantially the same general direction. The relative differences of phase between the resulting currents is utilized to select the desired current.

1,532,388—H. M. Dowsett, filed August 16, 1921, issued April 7, 1925. Assigned to Radio Corporation of America.

RADIO TELEGRAPH APPARATUS FOR AIRPLANES, consisting of an antenna system for aircraft where the bracing wires are located within the frame of the aircfaft and connected together and used as part of the radio signaling system.

1,532,523—F. Weinberg, filed August 11, 1923, issued April 7, 1925.

DETECTOR of the crystal type in which a pair of crystals are provided in vertical alignment with means for resiliently placing one crystal in contact with another. A threaded sleeve is provided upon the end of which is secured a crystal holder for maintaining the movable crystal in contact with the fixed crystal.

1,532,533—James E. Harris, filed December 16, 1919, issued April 7, 1925. Assigned to Western Electric Company, Incorporated, New York.

COLLOIDAL SUSPENSION for coating composition for filaments used in electron discharge tubes. The colloidal composition consists of particles of salt of an alkaline earth metal in solution. This colloidal solution is coated upon the cathode forming a thermionically active coating.

1,532,846—C. H. Thordarson, filed March 31, 1920, issued April 7, 1925.

ELECTRICAL CONDENSER designed for high tension operation, in which the conducting elements are formed of tube-like structures having side members for contact with the dielectric elements which are interposed between the tube-like conducting plates. The plates are so spaced that a cooling medium may be circulated therebetween.

1,532,964—C. Valguarnera, filed August 9, 1924, issued April 7, 1925.

MEANS FOR RADIO SIGNALING SHIPS DURING A FOG as a supplement to acoustic and luminous fog signals. The ships are equipped with special transmitting and receiving apparatus having limited range, each tuned to each other on a predetermined wave length. Signals are transmitted automatically and actuate the receiving station which is adjusted to receive the particular signals transmitted. As soon as the signals are picked up audibly a radio direction finder may be connected to the receiving set for accurately determining the position of the transmitting station.

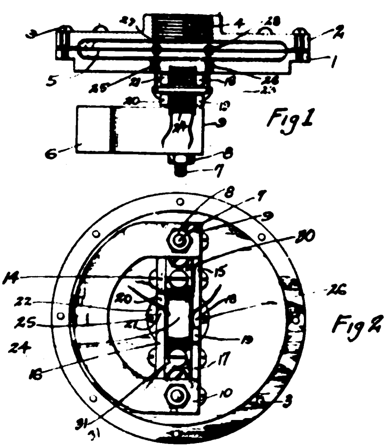
1,533,070—T. E. Arundel, filed May 26, 1924, issued April 7, 1925. Assigned one-fourth to Guy D. Shipherd and one-fourth to William H. Metcalfe, both of Omaha, Nebraska.

DETECTOR construction designed to maintain its adjustment permanently. The detector comprises a pair of electrodes which are secured within an insulated block and arranged to be adjusted exteriorly of the block and maintained in adjusted contact position for effecting rectification in an electrical circuit.

1,533,223—L. W. Chubb, filed June 30, 1921, issued April 14, 1925. Assigned to Westinghouse Electric and Manufacturing Company.

SYSTEM OF CONTROL for radio telegraph transmission wherein the signaling energy may be properly utilized between the signaling periods. A double antenna system is provided in which the antenna circuits may be employed as an absorbing circuit or as a radiating circuit and by varying the phase of the currents in the antenna circuits they may relatively add or subtract in their effects. By utilizing the phase control method the use of large absorbing circuits usually employed at an arc-signaling station is eliminated.

1,533,372—C. E. Brigham, filed June 14, 1924; issued April 14, 1925. Assigned to C. Brandes, Incorporated.

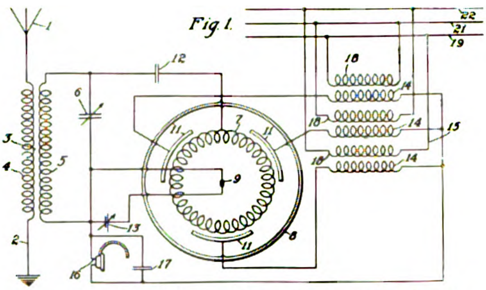


NUMBER 1,533,372—Loud Speaker

LOUD SPEAKER for radio reproduction, in which the maximum energy from the electromagnetic fluctuations is derived by means of a freely floating armature disposed in the magnetic field. The

armature is arranged within the electromagnetic field and is actuated by the variations in the magnetic flux in such manner that all the vibrations are effectively employed for the reproduction of signals.

1, 533,278—J. Slepian, filed November 20, 1920, issued April 14, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



Number 1,533,278—Plate Circuit Excitation

Plate Circuit Excitation for an electron tube system, in which multiphase alternating current is used for exciting the plate filament circuit. A polyphase source of electromotive force is included in the plate filament circuit to produce a flow of current therein similar to that produced by the direct current electromotive force method of plate excitation. A rotating electrostatic field is thus produced within a closed vessel, which field may be controlled for further controlling energy in the output circuit of the electron tube.

1,533,334—H. O. Russell and C. L. Paulus, filed April 18, 1922 issued April 14, 1925.

Condenser of the fixed type in which a stack of mica sheets having metallic coatings thereon are positioned adjacent each other; the mica sheets each have a coating of material subject to the action of electrolysis, such as lead oxide. On the surface of the lead oxide coating an electrically deposited plate of copper is formed. The mica sheets may then be stacked one adjacent the other to build up the condenser and the plates maintained under pressure in the stack.

1,533,502—J. L. Jenks, Jr., filed February 11, 1922, issued April 14, 1925.

Method of and Means for Adjusting Rotary Spark Gaps of Radio Apparatus by controlling the phase of the current supplied to the driving motor and thereby regulating the discharge according to the spark gap. The supply circuit is adjusted by means of variable reactance for controlling the phase of the energy to the motor. By a change in reactance sufficient lag may be introduced to obtain a consequent lag in the spark gap.

1,533,611—W. R. Respess, filed December 22, 1923, issued April 14, 1925. Assigned to New Jersey Research Company.

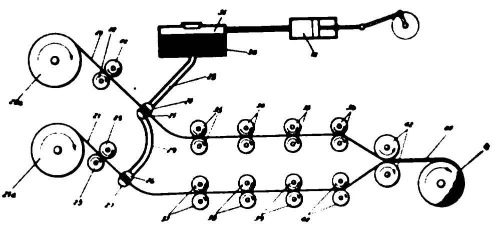
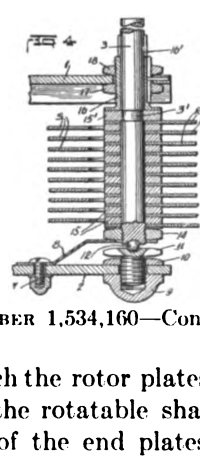


FIG 6

Number 1,533,611—Electrical Condenser and Method of Manufacturing the Same

Electrical Condenser and Method of Manufacturing the Same, where the metallic armatures are coated with rubber in solution which, when dry, forms an insulating filament directly upon the conducting plate, permitting condensers to be built up by stacking the conducting plates one upon another.

1,534,160—S. Cohen, filed February 17, 1925, issued April 21, 1925.



Number 1,534,160—Condenser

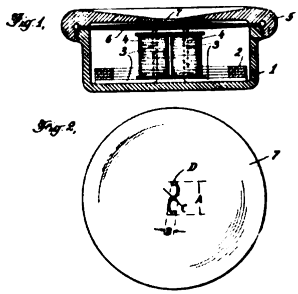
Condenser, in which the rotor plates are mounted on a ball-bearing member with the rotatable shaft journaled in a spring chuck secured in one of the end plates forming the condenser

frame. The condenser is designed for high electrical efficiency and high dielectric characteristics. The frame of the condenser and the rotor plates is of the same electrical potential.

1,534,213—G. Hill, filed February 3, 1923, issued April 21, 1925.

VARIABLE CONDENSER, in which the stator plates are cut from circular sheet material having a peripheral rim which extends around the plate in a complete circular form with a cut-out portion in the plate. The movable plates are arranged on a shaft and may be inter-leaved between the stator plates passing through the open portion of the stator plates. The stator plates are supported at different points around the edge thereof.

1,534,373—H. Fischer, filed October 25, 1922, issued April 21, 1925. Assigned to C. Brandes, Incorporated, New York.



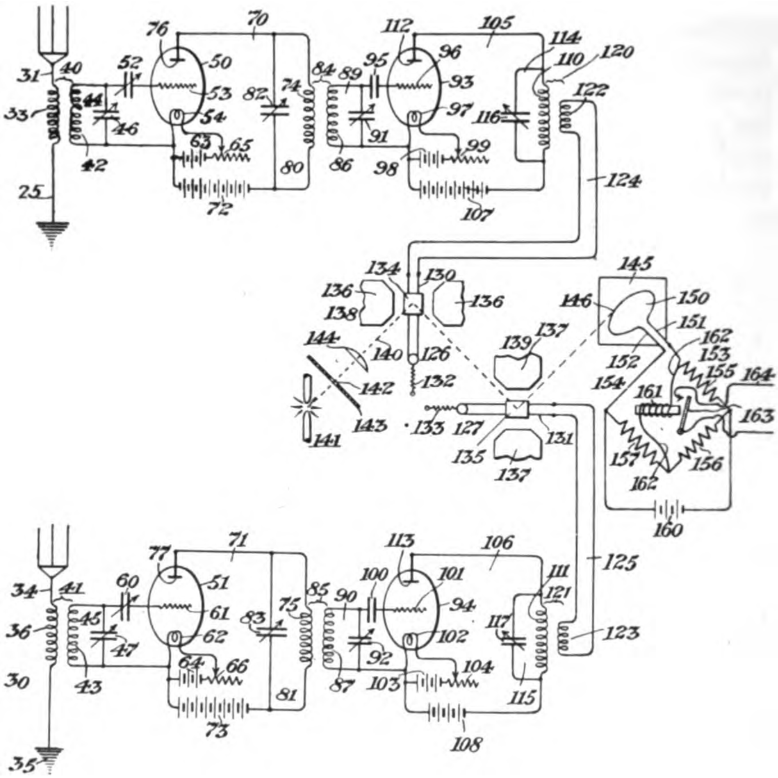
NUMBER 1,534,373—Diaphragm for Telephone Receivers

DIAPHRAGM FOR TELEPHONE RECEIVERS, consisting of a pair of super-imposed members forming an air chamber therebetween in which one of the members has an F-shaped slot therein. The object of the invention is to provide a diaphragm structure which will faithfully reproduce all of the tones of the musical scale. The patent covers an electromagnetic sound-reproducing mechanism having a base made up of a plurality of stamped metal parts. The invention is particularly adapted for quantity production of electromagnetic sound reproducers.

1,534,704—J. H. Hammond, Jr., filed September 9, 1918, issued April 21, 1925.

RECEIVING SYSTEM FOR RADIANT ENERGY, having separate circuits in which two series of impulses of radiant energy having

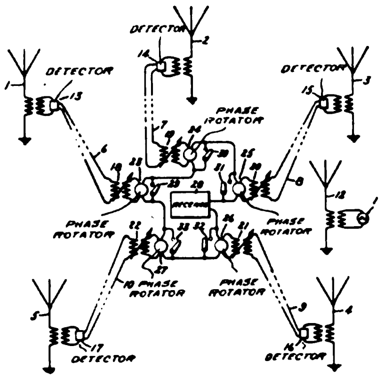
a predetermined phase difference may be utilized to control the operation of selenium cells at a receiver which in turn control circuits at the receiving station to selectivively actuate the receiving mechanism.



Number 1,534,704—Receiving System for Radiant Energy

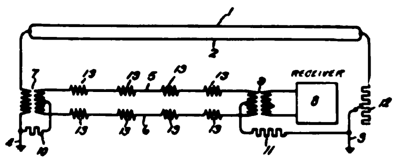
1,534,719—E. W. Kellogg, filed April 21, 1921, issued April 21, 1925. Assigned to General Electric Company.

Radio Receiving System, in which a plurality of widely separated receiving antennas are connected by transmission lines to a central receiving station. A radio frequency wave is radiated at a frequency differing from that of the signal waves to be received by an amount sufficient to produce audible beats with the signaling waves. At each receiving station the signaling wave is detected and the resultant audio frequency current is sent to the central receiving station where the signals may be received with the most favorable stray ratio. The multiple reception method insures the more accurate copying of the signals despite stray interferences.



NUMBER 1,534,719—Radio
Receiving System

1,534,720—E. W. Kellogg and C. W. Rice, filed May 18, 1921, issued April 21, 1925. Assigned to General Electric Company, New York.



NUMBER 1,534,720—Radio
Receiving System

RADIO RECEIVING SYSTEM employing a uni-directional horizontal receiving antenna for the reception of desired signals free of interference from disturbing waves coming from another direction. A long horizontal receiving antenna is provided with a transmission line running parallel with the antenna for conveying signaling currents from a selected point in the antenna to a distant receiving station.

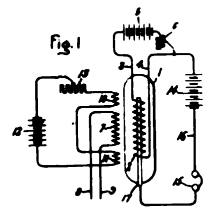
1,534,769—R. Brownlie, filed January 19, 1924, issued April 21, 1925.

OSCILLATION DETECTOR of the crystal type wherein the crystal is supported within a holder and a contact member arranged to be contacted thereagainst and moved by rotative movement to a desired point with respect to the surface of the crystal.

1,535,046—O. Scheller and R. Herzog, filed August 31, 1921, issued April 21, 1925. Assigned to C. Lorenz Aktiengesellschaft, of Lorenzweg, Berlin-Tempelhof, Germany.

Key Connection for Radio Transmission of Messages, for maintaining a steady load on a transmitting System. A circuit is provided at the transmitter where a choke coil may be brought into shunt relation to the antenna generator circuit when the transmitting key is closed, while a condenser is connected in parallel with the generator circuit when the key is open for alternately radiating and suppressing the signal energy.

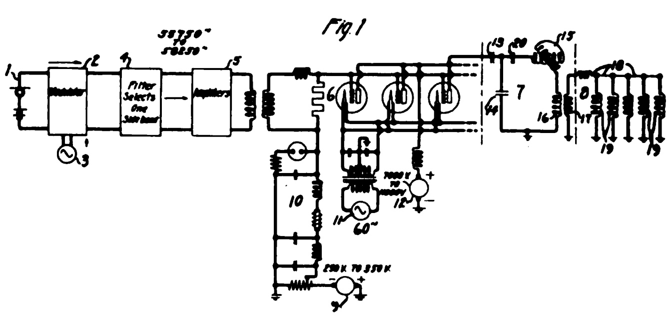
1,535,082—E. F. W. Alexanderson, filed September 28, 1920; issued April 21, 1925. Assigned to General Electric Company.



Number 1,535,082—Electron Discharge Device

Electron Discharge Device, in which an elongated anode is disposed within an evacuated envelope with a filamentary cathode symmetrically spaced around the anode. A magnetic field is generated substantially parallel to the axis of the cathode. By controlling the magnetic field the electron emission from the cathode is controlled to effect variation between the anode and cathode circuit.

1,535,130—A. A. Oswald, filed March 20, 1924, issued April 28, 1925. Assigned to Western Electric Company, New York.



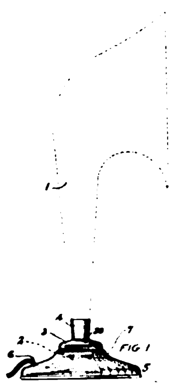
Number 1,535,130—High Power Radio Telephony

High Power Radio Telephony for long distance communication where a band of frequencies of the order of 2,000 cycles is employed. The antenna which is utilized has a resonance charccteristically materially narrower than 2,000 cycles with a substantially uniform effectiveness over the entire frequency range of the band.

1,535,189—R. E. Thompson, filed March 7, 1919, issued April 28, 1925. Assigned to Wireless Improvement Company, of Jersey City, a corporation of New York.

Radio Communication Apparatus, in which a single control is employed for tuning the apparatus. The variable inductance and condenser are mechanically adjustable thru the same interconnecting linkage to permit variation of the wave length of the circuit over a wide range by a single operation.

1,535,734—D. H. Moss, filed February 9, 1924, issued April 28, 1925. Assigned to C. Brandes, Incorporated, New York.



Number 1,535,734—Support for "Table Talkers" and Method of Making the Same

Support for "Table Talkers" and Method of Making the Same. The patent shows a method of construction for an electromagnetic sound reproducer. The construction of the base and means for mounting the acoustic reproducer therein is described in connection with the process of manufacture by which the instruments can be produced on a quantity-production scale inexpensively.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 13 AUGUST, 1925 Number 4

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GENERAL INFORMATION

The Proceedings of the Institute are published every two months and contain the papers and the discussions thereon as presented at the meetings in New York, Washington, Boston, Seattle, San Francisco, or Chicago.

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INSTITUTE ACTIVITIES

Standard Report

The Standards Committee of THE INSTITUTE, Ralph. H. Bown, chairman, is making good progress with the revision of the 1922 Standardization Report. It is expected that the work will be completed this Fall so that pamphlet publication of the Report may be possible early in the coming year.

Institute Sections

By authorization of the Board of Direction at its July, 1925, meeting, President Dellinger appointed a Sections Committee charged with the duty of investigating and developing the technical activities of present Sections, and of reporting upon by-law revisions which will enable the Board of Direction to administer more effectively the affairs of the Sections. The Committee is made up of Mr. Donald McNicol, chairman, and Messrs. Melville Eastham, J. V. L. Hogan, Leslie McMichael, H. M. Turner, G. Y. Allen, C. M. Jansky, Jr., J. C. Jensen, G. W. Pierce, A. H. Taylor, Montford Morrison, J. F. Dillon and C. E. Williams.

Increase in Membership

The Membership Committee, L. E. Whittemore, chairman, in a campaign carried on in June and July, last, succeeded in obtaining applications from about three hundred radio engineers and workers who desire to become associates of THE INSTITUTE. Thruout the Fall and Winter months the campaign shall be continued with a view to doubling the membership within the year.

Affiliation With A. A. A. S.

THE INSTITUTE recently became affiliated with the American Association for the Advancement of Science.

National Electric Code

THE INSTITUTE is now a member of the National Fire Protection Association, whose Electrical Committee originates the provisions of the National Electric Code. THE INSTITUTE OF RADIO ENGINEERS is represented on the Electrical Committee. Revisions of the Code are made and published annually. INSTI-

TUTE members should study the Section of the National Electric Code governing radio installations and forward to the Secretary any suggestions which would make for betterment of the wiring regulations covering radio and storage battery installation.

Chicago Section

The initial meeting of the Chicago Section of THE INSTITUTE was well attended and the first papers presented were of a high order. The speakers were Professor E. W. Bennett and Assistant Professor L. J. Peters, of the University of Wisconsin. Mr. Montford Morrison is chairman of the Section; L. R. Schmidt, secretary, and William W. Harper, treasurer.

New York Meetings

Growth in membership and in influence of THE INSTITUTE are reflected by the increasing interest being taken in the New York meetings. At recent meetings it has been necessary to use the large auditorium of the Engineering Societies Building in order properly to accommodate the audience. It is expected that in the near future attendance at meetings will exceed five hundred.

A NEW PHENOMENON IN SUNSET RADIO DIRECTION VARIATIONS*

By

L. W. Austin

(For Special Laboratory Radio Transmission Research)

(Conducted jointly by the Bureau of Standards and the American Section of the International Union of Scientific Radiotelegraphy).

The observations on the deviations preceding sunset[1] have been continued. The phenomena, it will be remembered, are as follows: The apparent direction of the long-wave stations, New Brunswick and Tuckerton, to the northeast of Washington, begins to shift toward the east two or three hours before sunset. This deviation reaches a maximum of 10° to 15° roughly an hour before sunset. The bearing returns to normal before sunset and then usually shifts to the west and passes into the irregular night deviations.

The remarkable thing about this phenomenon is its uniformity, the only variations from day to day being differences in the amount of deviation and the exact time when the bearing returns to its correct value. It seems to occur with regularity only with stations at certain distances, not over 300 km. and not less than 100 km. As the only stations suitable for these observations lie to the northeast of Washington, attempts have been made to interest observers in taking observations in other directions. Work covering only two or three days by Mr. Englund at Cliffwood, New Jersey, indicated that Annapolis, about 270 km. to the southwest, showed deviations first to the west and later to the east, that is, in the opposite sequence to those observed on the northerly stations at Washington.

According to Eckersley's theory,[2] deviation is due to an indirect wave reflected or refracted from the Kennelly-Heaviside layer, which comes down with its magnetic field non-parallel to the earth's surface, thus having a vertical component which cuts

*Published by permission of the Director of the Bureau of Standards of the United States Department of Commerce. Received by the Editor, June 12, 1925.

[1]Proceedings of The Institute of Radio Engineers, 13, page 3; 1925.
[2]"Radio Review," 2 page 60; 1921.

the top and bottom of the [illegible] an [illegible] which destroys the true [illegible] and requires turning the compass coil to bring the [illegible] balance. If we could assume that the reflecting layer is horizontal and that there is a regular reflection it should be possible to restore the bearing to the true direction by rotating the frame carrying the compass coil about a horizontal axis at right angles to the line joining the stations. Then at the vertical angle which restores the true horizontal bearing the compass coil should be at right angles to the direction of the incident wave, and from the assumptions should enable the height of the reflecting layer to be determined by triangulation. This experiment has been tried but failed to restore the true bearing. It did, however, show a new series of phenomena which, while not so far explained, show apparently the same regularity of sequence as the before-sunset deviation already described. It has been frequently noticed that the rotation of the normal axis of the compass coil around a horizontal axis at right angles to the true direction of the sending station frequently produces a great sharpness of minimum at a certain vertical angle. It is now found that the angle for the sharpening of the minimum apparently varies regularly with the changes in bearing deviation during the before-sunset period. The "sharp minimum" vertical angle starting at 0° to 20° increases with the deviation of the horizontal bearing until at about an hour before sunset, just before the horizontal bearing deviation has reached a maximum, it reaches 50° to 80°. Then, as the horizontal bearing returns toward the true direction, the vertical "sharp minimum" angle decreases rapidly so that before the horizontal bearing has become correct, the vertical angle has returned to zero and gone up to 50° to 80° on the other side, that is, with the main axis of the compass coil tipped forward.

Figure 1 shows the double axis compass coil and Figure 2 a typical set of curves. A few points in regard to the curves are worthy of notice. The sharp minimum vertical angle always begins to rise some time before the bearing of the station begins to shift. The vertical angle, so far as has been observed, always returns to zero at the same moment that the easterly bearing deviation begins to drop. The vertical angle curve cuts the axis again nearly at the same time that the westerly deviation starts to decrease. The negative maximum of the vertical angle always nearly coincides with the passage of the bearing thru its true value in going from the easterly to the westerly deviation.

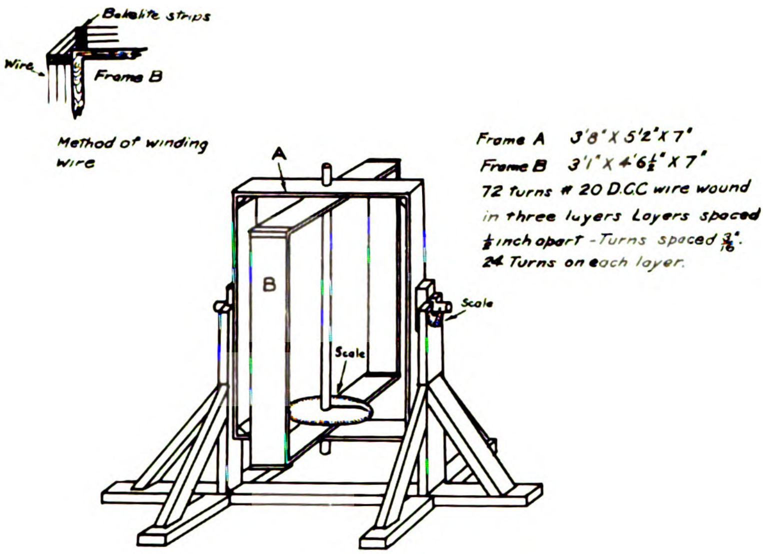

FIGURE 1

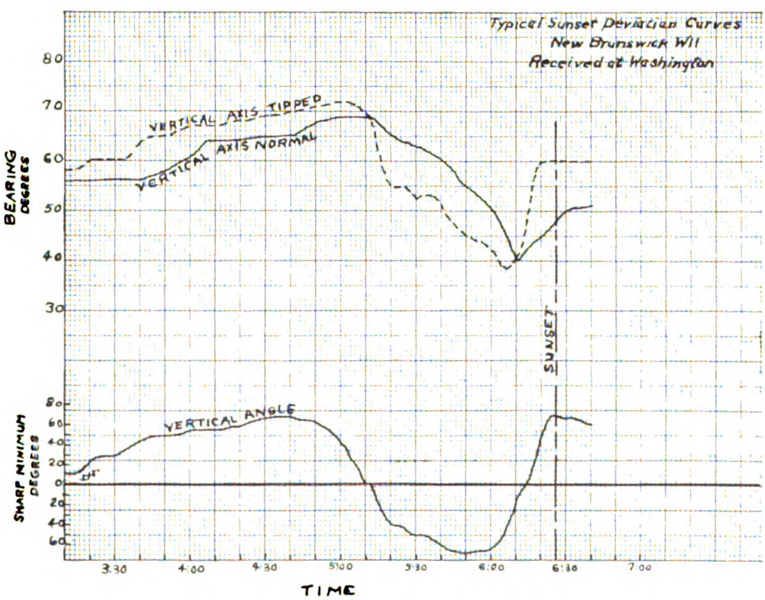

FIGURE 2

Enough observations have been made to convince us that we have a perfectly regular natural phenomenon apparently occurring daily, which is probably connected with the deionization of the atmosphere as the sun sinks toward the west. It seems probable

that similar deviations take place after sunrise, but these have not as yet been investigated. As it may be a long time before the physical processes involved are understood, I am publishing the observed facts for others to verify and, if possible, explain.

RECENT COMMERCIAL DEVELOPMENT IN SHORT WAVE TRANSMITTERS AND RECEIVERS*

By

S. E. Anderson, L. M. Clement, and G. C. De Coutouly

(Western Electric Company, New York)

Introduction

In view of the recent activity in the use of the very short wave lengths, that is, below 100 meters, it may seem somewhat tardy to present a paper describing a transmitter and receiver recently built for the United States Coast Guard covering the wave length range between 100 and 200 meters (1,500 to 3,000 kc.). When the complete title of the paper is taken into consideration, however, the authors hope that the matter presented will be of interest, as commercial development always follows somewhat behind the development of the laboratory.

The paper deals with the development of the radio receiver and transmitter for telephonic and interrupted continuous wave telegraph transmission for communication between small vessels and between them and shore stations. The sets must be designed to give satisfactory communication over distances of 50 miles for telephony and 100 miles for interrupted continuous wave telegraphy. One wave length only will be used for communication and this equipment will be operated by persons not familiar with radio. The controls on the receiver, therefore, will be locked and the transmitter will be adjusted to a single frequency. In order to insure absolutely reliable communication the selectivity of the receiver must be good and the transmitting frequency must be held within very close limits.

The system consists of two main parts—the 50-watt coupled oscillator circuit type of transmitter and the double detection type (super-heterodyne) receiver.

The primary power source is a 200 ampere-hour storage battery which supplies the filaments of the transmitter and the plate circuit dynamotor. A standard telegraph key is used to con-

* Presented before The Institute of Radio Engineers, New York, December 17, 1924. Received by the Editor, January 13, 1925.

trol the telegraph feature of the transmitter while a hand set is used for telephony. Separate dry or storage batteries are used to operate the radio receiver.

The Transmitter

The operating conditions demand that the transmitter be simple to operate, rugged, and efficient in order to insure minimum drain on the storage battery. The frequency stability requirements are very severe as it is necessary to maintain the transmitter frequency constant for variations in the supply voltages and change in antenna constants due to service conditions. The most serious cause of variation is due to changes in the antenna capacity caused by the rolling of the vessel, which may be as great as 45° from the vertical. The range requirements of 50 miles for telephone and 100 miles for telegraph signals are considered to be satisfactorily met by a 50-watt transmitter as a result of some preliminary tests made by the Coast Guard people. The power output requirements of the set expressed in terms of the antenna constants and antenna current demand that it deliver at least two amperes into a 12-ohm antenna of 0.0003 microfarad capacity having a natural wave length of 116 meters.

Electrical Design

The transmitter circuit is made up of three major parts—the oscillator circuit, the modulator circuit and the speech input circuit. The circuit diagram of the transmitter is shown in Figure 1. The negative grid biasing potentials for all tubes are obtained by means of a system of resistances properly inserted in the plate circuit. Keying for interrupted continuous wave telegraph transmission is obtained by the use of a 15,000-ohm resistance (R_{13} on the diagram) which is connected between the minus 18-volt terminal and the ground. The key short-circuits this resistance when it is depressed, which removes the high negative potential applied to the grids of the tubes.

The completed transmitter is shown pictorially in Figures 2, 3, and 4. It consists of a single unit brass frame 33″ high, 16½″ wide, and 18″ deep. The upper part of the front panel supports all of the meters which are necessary for observation of the performance of the different parts of the circuit. At the bottom of the front panel the switches for the control of the filaments and the motor-generator are located. The necessary connections to the set are made to a terminal strip which is accessible thru

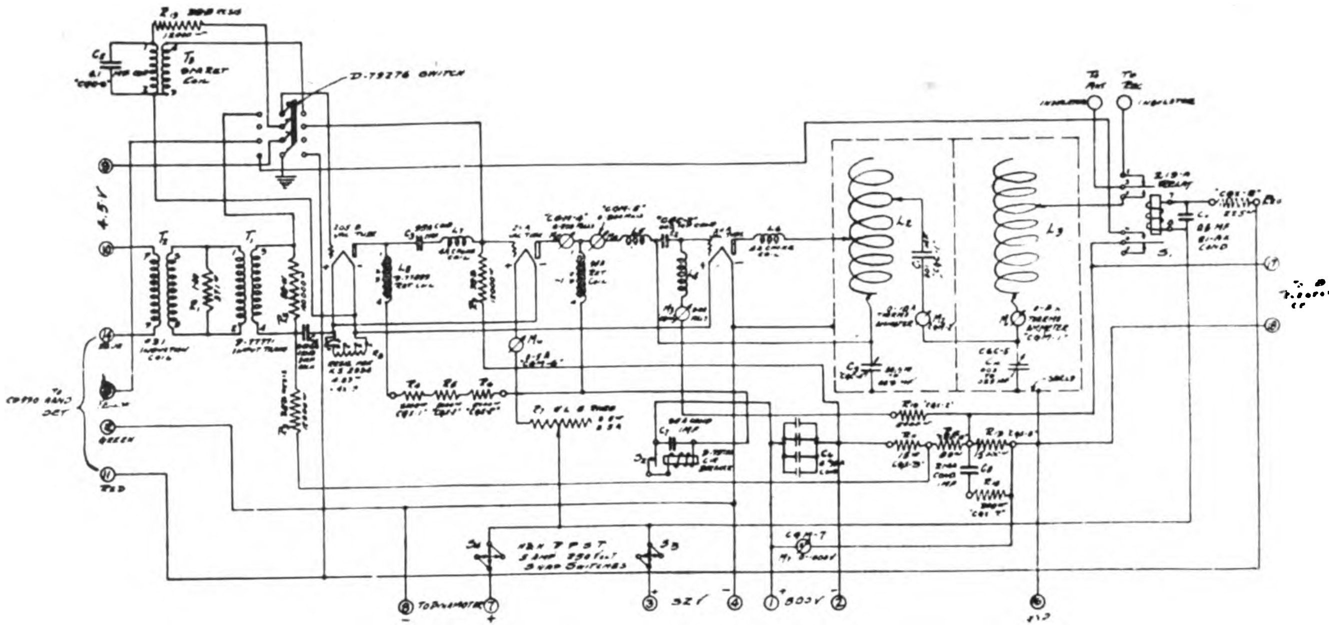

FIGURE 1—Schematic Circuit of Complete Transmitter

a door at the bottom of the transmitter. The equipment in the transmitter as shown in the photographs is mounted in three separate sections. In the first section from the bottom audio frequency coils and resistance units for the speech input and modulator circuit are mounted.

The second section from the bottom is divided into two compartments—the front containing the two power tubes as well as the speech input amplifier tube with their accessories. The back is entirely shielded from the rest and contains the primary tuning inductance, the primary tuning condenser and the grid coupling condensers. The third section located at the top of the set is also

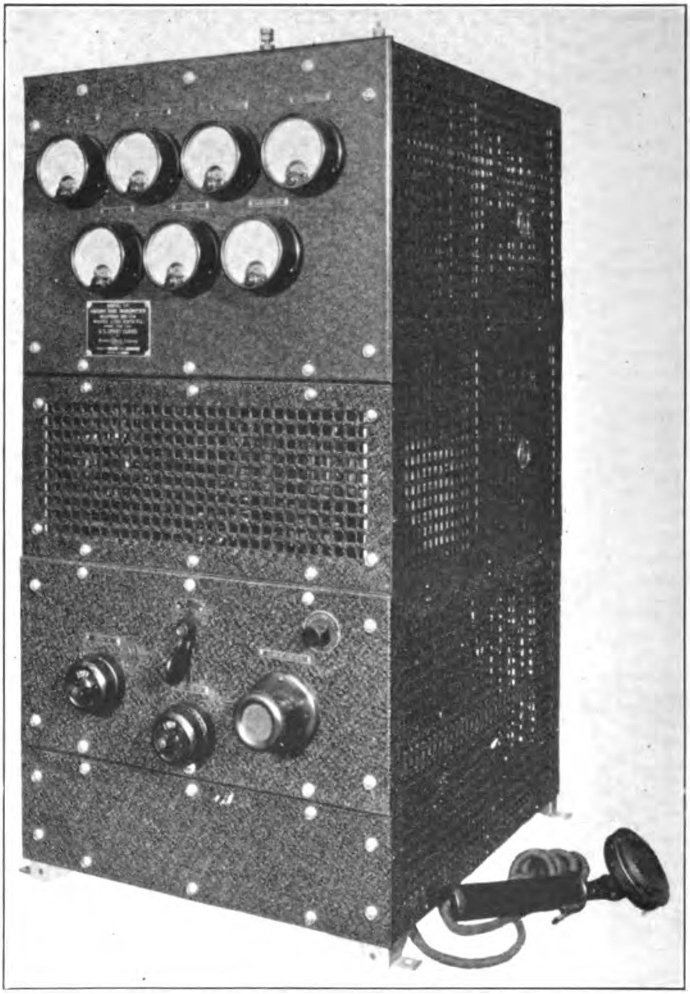

FIGURE 2

divided into two parts, the front containing the plate voltage meter resistor and the antenna relay and the second containing the antenna loading coil and the antenna coupling condensers. This compartment is also entirely shielded from

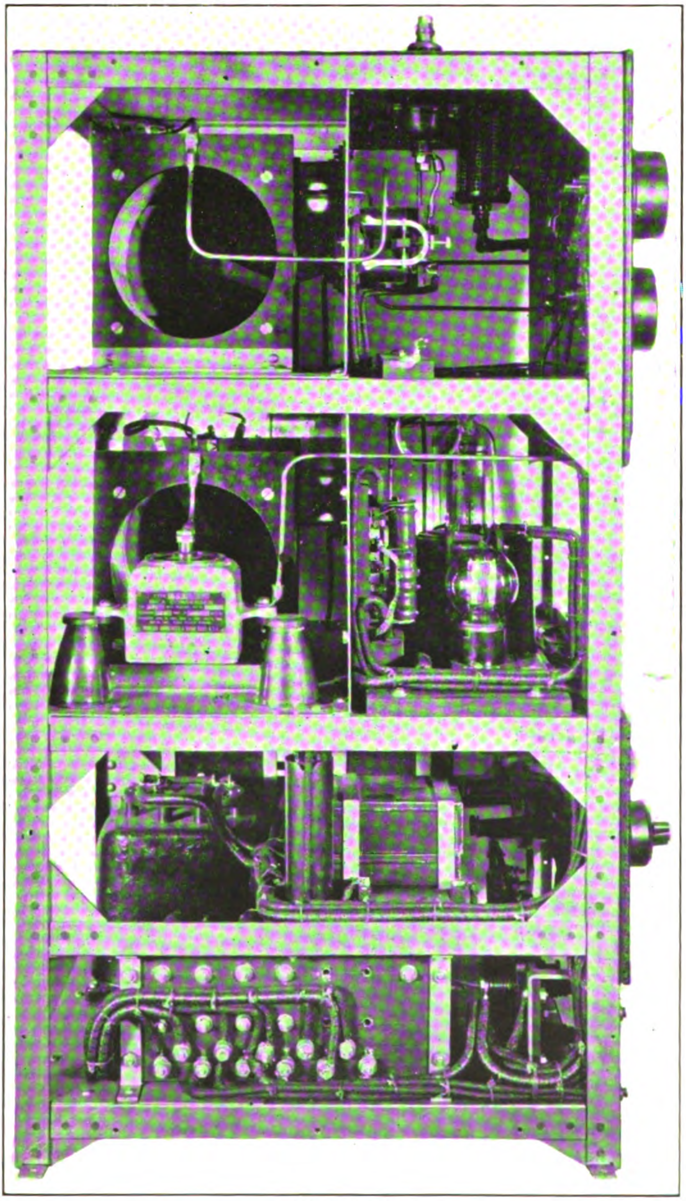

FIGURE 3

the rest of the circuit. The control switch for telephone or interrupted continuous wave telegraph transmission is mounted at the center of the front panel. When the set is operating as a telephone transmitter the push button provided on the hand set connects the transmitter to the antenna and

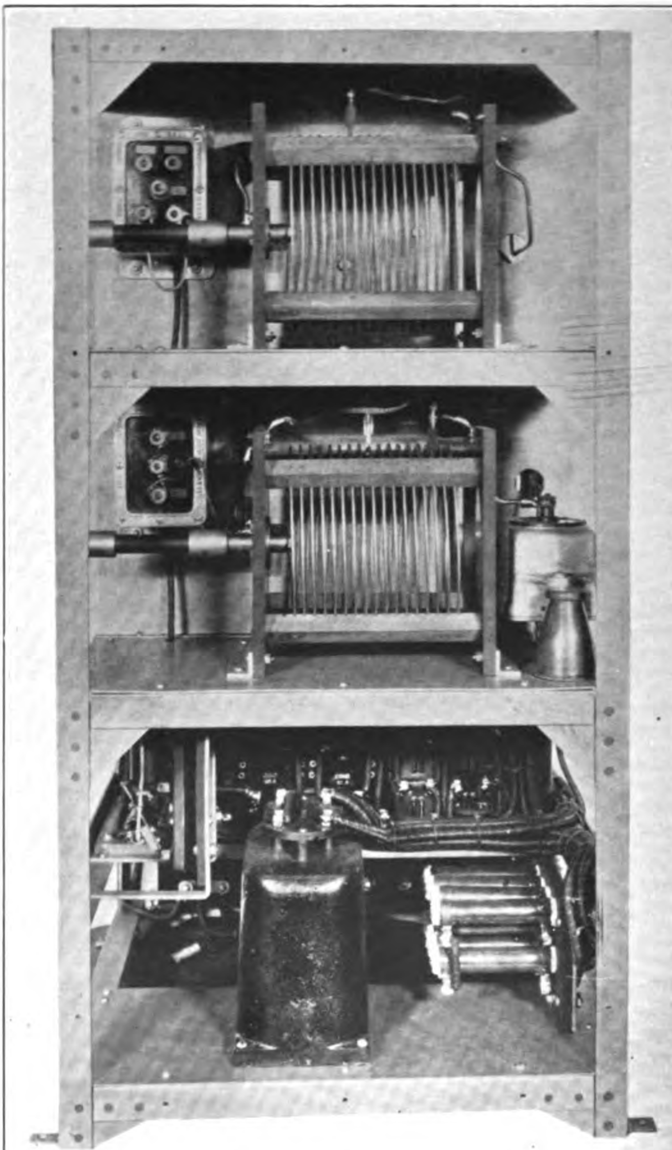

FIGURE 4

starts the oscillations when it is pressed. With the control switch in the interrupted continuous wave position the telegraph key is connected in circuit and oscillations are started when the key is pressed. The filaments of the tubes are controlled by means of an ordinary snap switch which is shown on the photograph, Figure 2. A second snap switch controls the plate dynamotor which supplies the plate voltage to the plates of the tubes only after the filaments of the tubes have been turned on. A filament rheostat is also provided to adjust the filament current to the proper value.

The antenna and primary coils are provided with a fine continuous adjustment of their inductance so that the circuits may be adjusted to any frequency within the range of the transmitter. The fine adjustment is obtained by means of a sliding contact on the last turn of the inductance. It can be set in any position from the outside of the transmitter by means of a screwdriver. No fine adjustment knobs were provided, as the transmitter is to be set at the desired frequency, and it will remain constant for a long period of time.

One of the first problems to be considered was the choice of the type of transmitter circuit which would best give the degree of frequency stability required. With this end in view, both the simple oscillator circuit and the coupled oscillator circuit were studied.

The change of antenna capacity for a 45° roll of the vessel from the vertical was calculated by well-known formulas and found to be approximately 3 percent of the total antenna capacity. The frequency change due to a 3 percent variation in antenna capacity was then computed for the simple tuned oscillator circuit and for the capacity coupled oscillator circuit.

The calculated frequency change for the case of the simple oscillator for a 3 percent increase in antenna capacity was found to be approximately 8,000 cycles at an operating frequency of 2,300 kilocycles.

In the case of the capacity coupled oscillator circuit the circuit constants were determined for the resonance condition of the double tuned capacity circuit with an antenna of 0.003 microfarad capacity. These constants were calculated so that the circuit resonated at the desired frequency F_1. Assuming that the change of frequency due to the antenna capacity variation was of the second order with respect to the change of frequency in the primary circuit due to the change of the introduced reactance from the secondary into the primary, the reactance introduced

the secondary into the primary after the antenna capacity change was calculated as if the circuit was still resonating at the original frequency F_1.

The frequency at which the primary circuit resonates after the antenna capacity change was obtained by calculating the frequency determined by the primary inductive reactance in series with the reactance of the grid and primary tuning capacities and the value of reactance introduced from the secondary circuit after the antenna capacity change. This gives a new value F_2 for the frequency which was then substituted back into the equations for the secondary circuit in order to calculate more accurately the change of apparent reactance introduced into the primary circuit by the secondary. By means of several approximations the final change of frequency is approached, and the difference between it and the original frequency F_1 gives the change due to the changed antenna capacity. The frequency change was found, in this case, to be about 3,100 cycles per second when the coupling between the primary and secondary circuits was of the critical value (critical coupling).

The simple type oscillator did not meet the frequency stability requirements, and accordingly the capacity coupled oscillator circuit was chosen. The coupled oscillator circuit does not radiate harmonic frequencies of the fundamental to any great extent, and this alone is enough to justify its use.

The laboratory model and later the completed commercial transmitters were tested for frequency stability due to changes in plate potential supply voltage and change in antenna capacity.

As the change of frequency due to change in the antenna capacity and to the variations of plate and filament voltages due to the discharge of the battery was very small in comparison with the frequency at which the transmitter operated, the following method for measuring this frequency change was used: The continuous wave emitted by the transmitter placed in the telephone position by means of the transfer switch, not modulated by the voice, was received with a heterodyne receiver which was set to give a certain beat frequency for a given condition of the transmitter. The frequency of this beat note was measured by means of an auxiliary standard calibrated oscillator by the "zero beat" method. The setting of the transmitter was then changed to simulate the changed condition, and the frequency of the heterodyne beat note in the receiver was again measured by means of the calibrated audio frequency oscillator. The difference between the two readings of that oscillator gave the difference

in frequency between the two conditions of the transmitter. The following results were observed:

For a variation of plate potential caused by a change in storage battery voltage from 33 (fully charged) to 28 volts (nearly discharged) with the filament kept at a constant value, the frequency change was of the order of 0.007 percent. It may, therefore, be said that the frequency of the transmitter is practically independent of the variations of plate voltage and storage battery voltage likely to occur in practice. The same method was used to measure the change of frequency due to a change of antenna capacity. For an antenna capacity change of from 300 to 310 micro-micofarad (3 percent) the change of frequency was found to be about 2,500 cycles per second when the antenna coupling capacity was chosen so that the required antenna power was obtained. During the test the antenna capacity was varied from 300 micro-microfarads to about 500 micro-microfarads. It was noticed that the frequency of the transmitter first increased with the antenna capacity and then decreased as the antenna capacity was further increased and approached a value after which no further decrease occurs. This series of observations at first could not be easily explained, but upon a mathematical investigation it was found to follow the theory.

Design of Radio Frequency Oscillator Circuit

The double-tuned self-excited oscillator output circuit is shown in Figure 1. It consists of a closed inductance-capacity oscillating circuit coupled to the antenna circuit by means of a large capacity which is common to both circuits. L_2 and C_1 are the primary tuning inductance and capacity, respectively, with L_2 variable by taps. C_{10} is the antenna coupling condenser and C_9 the grid coupling condenser. The circuit is grounded at the point common to condensers C_9 and C_{10}. The direct current plate voltage is applied to the oscillating tube thru the primary inductance L_2 in order to eliminate a radio frequency plate feed choke coil. The power delivered by the vacuum tube to the primary circuit is controlled by means o' the plate tap. The antenna circuit is composed of an antenna loading coil L_3, variable by taps, and in series with the antenna ammeter and the antenna coupling condenser C_{10}.

The values of the various circuit constants of the Coast Guard transmitter were calculated in two steps from the knowledge of the antenna and tube characteristics and the $\omega L/R$ assumed from previous experimental data. First, by means of

primary calculations the tuning and coupling capacity to be used in the primary closed oscillating circuit was determined from considerations of tube impedance and frequency. The second step of the caluclations was based upon the knowledge of the antenna coupling and primary capacity which had been obtained in the preliminary calculations. The values of the primary and secondary inductance were calculated as well as the effective resistances of the primary and secondary circuits and the currents and voltages in all branches of the circuits.

BASIS OF ALL CALCULATIONS

The resonance conditions that hold in practice for the calculation of double tuned capacity coupled circuits are:

$$1\text{st}—X_3=0$$

$$2\text{nd}—X_1'=0$$

X_3 being the reactance of the secondary circuit considered as a separate unit from the primary and X_1' being the apparent reactance opposed by the entire double-tuned circuit to the vacuum tube.

It is known that the plate-to-filament impedance of any vacuum tube is essentially a pure resistance at the frequency considered, and for this reason the load circuit should be adjusted to look like a pure resistance for best efficiency. This explains why the condition $X_1'=0$ has to be considered as one of the resonance conditions for that type of circuit. If these two conditions obtain, the reactance introduced into the primary by the secondary is equal to zero and the resistance introduced into the primary is equal to

$$R'=\frac{X_c^2}{R_3}$$

where X_c is the reactance of the coupling condenser at the frequency considered and R_3 is the effective resistance of the entire secondary circuit, that is to say, the sum of the antenna resistance proper and the resistance of the antenna loading coil at the operating frequency. The resonance conditions $X_1'=0$ may be written in the following form:

$$X_1'=X_1-\frac{X_1^2X_2'}{Z_2'^2}$$

where X_1 is the coupling reactance between the tube and the primary circuit, X_2' the apparent reactance of the secondary, and Z_2' the apparent impedance of the secondary.

However, $$X_2' = X_2 - \frac{X_c^2 X_3}{Z_3^2}$$

and as $X_3 = 0$, X_2' is equal to X_2. The resonance relation $X_1' = 0$ may be therefore written

$$X_1 - \frac{X_1^2 X_2}{R_2'^2 + X_2^2} = 0$$

This is equivalent to $X_2^2 = X_1 X_2 + R_2'^2 = 0$, as X_1 can never be equal to zero. This last equation in connection with equation $X_3 = 0$ enables all the circuit constants to be calculated.

As a first approximation in the calculations X_2 may be considered as strictly equal to "0" as its value.

$$X_2 = \frac{1}{2}\left(X_1 - \sqrt{X_1^2 - 4 R_2'^2}\right)$$

is so small that it can be neglected. The current amplitude in the secondary circuit may, therefore, be calculated by the formula

$$I_3 = \frac{X_1 X_c E}{R_1 R_2' R_1'}$$

where X_1 is the reactance of that portion of the primary inductance coil between the low point of the coil and the plate tap at the operating frequency. E is the amplitude of the effective driving electromotive force. R_2' is the apparent resistance of the secondary circuit:—

$$\left(R_2' = R_2 + \frac{X_c^2}{R_3}\right)$$

and R'_1 is the apparent resistance opposed to the vacuum tube of the entire double tuned capacity coupled circuit:—

$$\left(R_1' = R_1 + \frac{X_1^2}{R_2'}\right)$$

where R_1 is the impedance of the plate-to-filament circuit of the vacuum tube for alternating currents. This impedance is sufficiently independent of the frequency, so that it may be taken as a basis of calculations.

In order that the double tuned capacity coupled oscillating circuit be stable in operation, the coupling between the primary and the secondary should be very loose. If the coupling is tight, there are two frequencies at which this kind of circuit may oscillate, one of which is determined by the primary and the other by the secondary. The circuit should always be operated at or near the critical coupling value, for which the relation between the coupling reactance and the primary and secondary effective resistance is $X_c^2 = R_2 R_3$.

Experimental Development of the Transmitter

An experimental model of the transmitter was set up in the laboratory taking care that the relative position of all apparatus, the constants of which were calculated in accordance with the method described, was such as to simulate as nearly as possible the exact dispositions of apparatus chosen for the final transmitter. The circuit constants were adjusted to the values which had been determined by the theoretical design. The theoretical calculations were, in general, fully verified by measurement except for a few minor modifications which had to be made in the values of the antenna coupling capacity. These minor modifications were due to the fact that the coils utilized had more resistance than correspond to the assumed $\omega L/R$ in the theoretical calculations. This increase in resistance was caused by the additional losses due to the proximity of the shields to the coils. The performance of the audio frequency speech input amplifier and oscillator circuit was found to be in accordance with the theoretical considerations.

Audio Frequency Characteristics of the Transmitter and Percentage of Modulation

The frequency characteristic of the transmitter was determined by feeding into the input of the set at the two terminals where the microphone transmitter is connected a number of audio frequency currents at a level corresponding to the energy level furnished by the microphone transmitter, the transmitter being adjusted for telephone transmission. The value of the audio frequency input current was chosen so that no current was taken by the grid of the modulator tube. The output of the modulator tube was measured for each input frequency by means of a peak voltmeter placed across the modulator plate choke coil. The frequency characteristic is shown in Figure 5. It is seen that this characteristic is practically flat between 100 cycles and 2,000 cycles and does not vary widely over a much larger range. It is, therefore, evident that the quality of audio frequency transmission of the Coast Guard transmitter is excellent for practical telephone conversation.

Percentage of Modulation

The percentage of modulation was determined by an oscillograph in the ordinary way. An oscillogram was taken of the rectified wave when the transmitter was set for telephony and interrupted continuous wave. The two oscillograms, **Figure 6,**

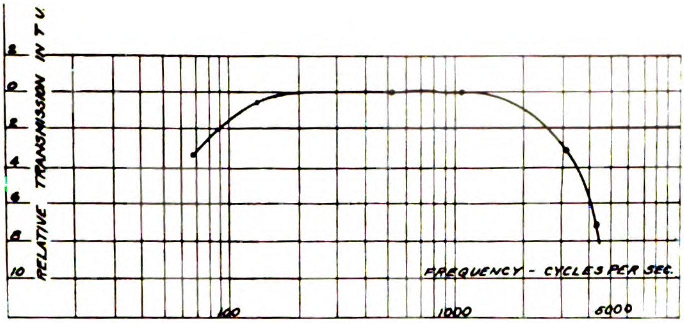

FIGURE 5—Frequency Characteristic Coast Guard Transmitter Model T-1

show that the percentage of modulation was 41 percent at an input audio frequency of 264 cycles and that the modulation of the carrier wave for the interrupted continuous wave telegraph position was complete. The effect of inertia of the moving element probably accounts for the peaks which appear below the zero line. The tone received, however, is very clear.

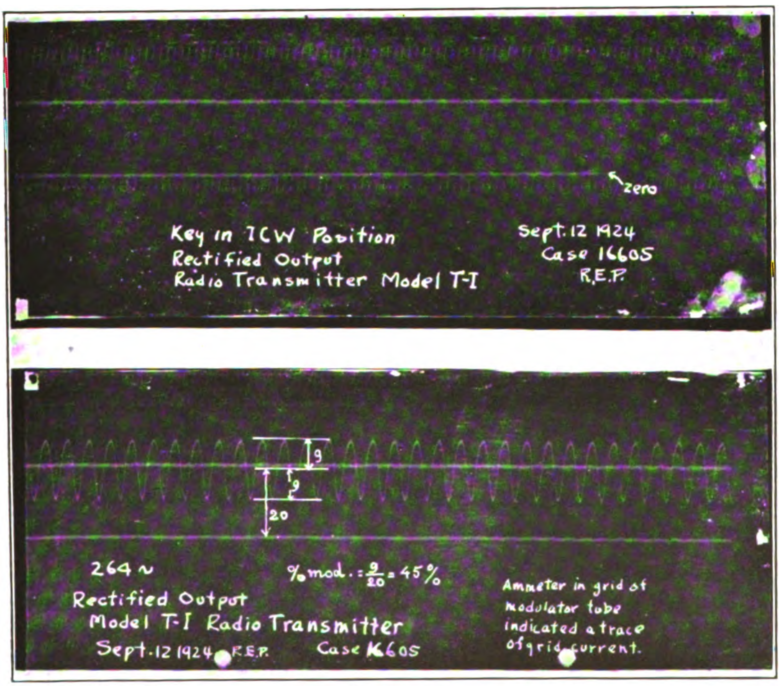

FIGURE 6

RADIO RECEIVER

The double detection (super-heterodyne) type of receiver is used because the required sensitivity and selectivity could be obtained only with this type of receiver. It is designed to cover a frequency range of 1,500 to 3,000 kc. (200 to 100 meters) and for the reception of telephone and interrupted continuous wave telegraph signals. In addition, a second oscillator operating at the intermediate frequency is provided for the reception of continuous wave telegraph signals. The development of this receiver involved the solution of a number of difficult problems in order to meet the rigid requirements imposed. In brief, it had to be sensitive (voltage amplification in excess of 5,000), selective against signals differing widely from the desired transmitter frequency but capable of receiving signals when the carrier frequency did differ from time to time not more than 5 kc. from the specified frequency. Its adjustment had to be simple and locks had to be provided so that it could be set and placed in operation by the turning of only the filament switch. Figures 8, 9, and 10 show the arrangement of the apparatus in the receiver and its general appearance. It is thoroly shielded by the brass panel shelf and the shielded box.

The complete receiver circuit (Figure 7) may be divided into the following parts:

Radio Frequency Input Circuit.
Radio Frequency Oscillator Circuit.
Modulator or First Detector Circuit.
Intermediate Frequency Amplifier Circuit.
Detector (Second) and Audio Frequency Circuits.
Intermediate Frequency Oscillator Circuit.

CHOICE OF INTERMEDIATE FREQUENCY

Taking into consideration the fact that the intermediate frequency selectivity is of no value in differentiating between two signals, the carrier frequencies of which differ by twice the intermediate frequency, a moderately high intermediate frequency would naturally be chosen. With too high an intermediate frequency the amplification obtainable is considerably reduced and the regenerative effects due to interstage coupling of one form or another are greatly increased. If too low a frequency is used, the tuning of the secondary circuit and that of the oscillator will differ by only a small percentage of the carrier frequency, and the tuning of the two circuits will not be independent of each other.

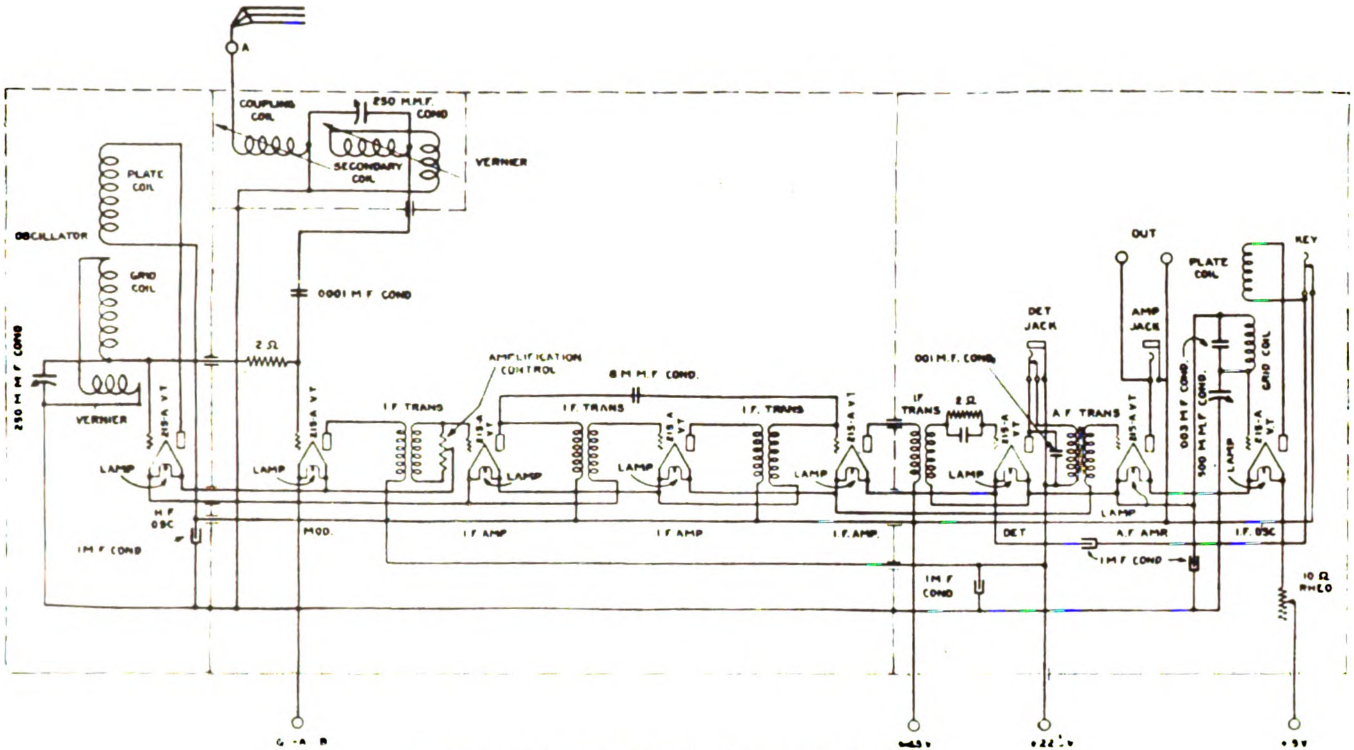

FIGURE 7—Schematic Circuit of Complete Receiver

The 50 kc. frequency used was chosen because satisfactory intermediate frequency transformers had been developed for this frequency, the required amplification could readily be obtained, and such an amplifier requires no stabilizing adjustment in order to prevent a tendency toward internal oscillation.

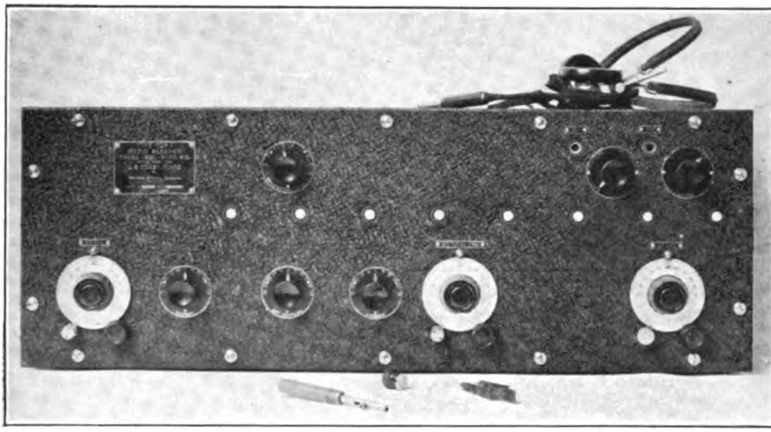

FIGURE 8

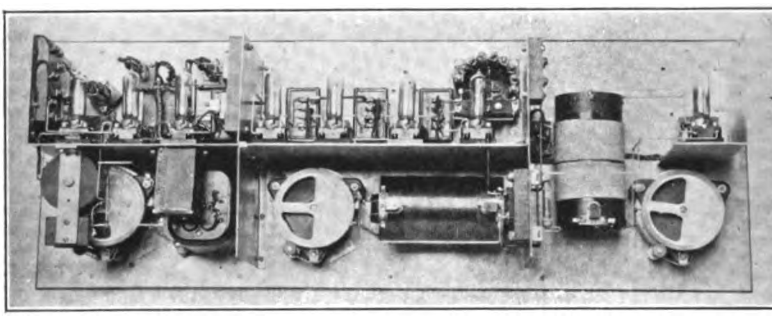

FIGURE 9

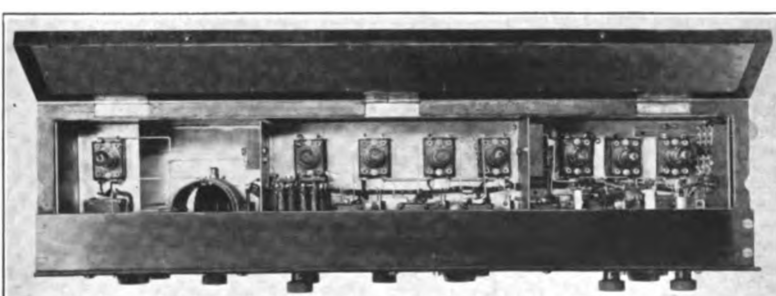

FIGURE 10

Radio Frequency Input Circuit

The radio receiver is designed to operate in connection with an antenna, but it is desired not to tune the antenna circuit in order to eliminate the additional tuning control. The radio frequency circuit, therefore, is of the simplest possible type, as shown in the schematic of the complete receiver. A small coupling coil is connected directly between the antenna and ground, the mutual inductance between this coil and the secondary circuit being adjustable. On account of the fact that if more than critical coupling is used, no additional signal strength is obtained and the selectivity of the secondary circuit is seriously impaired, the coupling coil was made of such size that the maximum coupline between the antenna and the secondary was not greatly in excess of critical coupling. The coupling coil is mounted at the low potential end of the secondary coil, and as one end of each is connected to the ground the capacity coupling between them is reduced to a minimum. The secondary coil is wound with bare copper wire spaced by its own diameter on a thin walled tube. This insures a coil of very low radio frequency resistance and, in fact, this is found to be practically as low as that obtained with the best types of cellular windings without any supporting dielectric. A "vernier" adjustment is provided by means of a small coil mounted at the opposite end of the secondary from the antenna coupling coil, but connected in the low potential side of the tuned circuit in order to have one terminal of the "vernier" coil at ground potential. The use of the inductance "vernier" has many advantages over the use of a separate plate on the variable condenser. With the latter the capacity of the separate plate may be 1/10 or 1/20 of the total capacity of the condenser, but when the condenser is adjusted to some point near its minimum capacity, the capacity of the "vernier" may be considerably greater than that of the condenser and its adjustment correspondingly critical. With the inductance "vernier," on the other hand, the percentage change in inductance and in the resonant frequency is nearly constant over the entire range of the receiver. The "vernier" was so designed that the total variation was equivalent to about 10 percent of the condenser setting for the major portion of its range. The adjustment of a "vernier" of this type is no more critical at the higher frequency end of the receiver range than at the lower.

Radio Frequency Oscillator Circuit

The tuned grid inductively coupled oscillator circuit was used

so that the tuning condenser is placed across only the grid coil. One side of it is thus at filament or ground potential with consequent elimination of the effect of the hand on the frequency of the oscillator. The inductance "vernier" is used in series with the grid coil and is mounted at the high potential end of the coil. The oscillating coils are so designed that the reading of the scale of the oscillator condenser for a frequency 50 kc. below that of the incoming signal is practically the same as the scale reading of the secondary circuit of the receiver over the entire tuning range.

Modulator or First Detector Circuit

Because the grid condenser and grid leak type of modulator requires a much smaller input on the grid for maximum efficiency than the negative grid bias type of detector, it is employed in this receiver. The circuits used for the frequency changing system consisting of the oscillator and modulator are illustrated in the schematic of the complete receiver, Figure 7. The condenser and grid leak combination is chosen for maximum efficiency, which is obtained with a capacity of 100 micro-microfarads and a 2-megohm resistance leak.

The grid leak serves both as a grid leak for the modulator tube and as a means of coupling the modulator tube to the oscillator. With this circuit the adjustment of the secondary circuit has almost no effect on the frequency of the oscillator except when the secondary circuit is made resonant with the oscillator frequency, which is not the operating condition. The oscillator is carefully shielded from the remaining portion of the receiver in order to prevent any interaction between it and the secondary circuit.

Filament Circuit

The vacuum tubes used in this receiver each require a filament current of 250 milliamperes at approximately 1 volt. When employing a number of tubes of this particular type it is advantageous to connect their filaments in series, so that grid biasing potentials may be obtained by the drop in potential across certain portions of the filament circuit. In a receiver having high amplification, connecting the filaments in series presents some additional problems because of the coupling thus introduced between the grid circuits of the various tubes. This coupling is reduced to a satisfactory point in this receiver by the use of a number of high capacity by-pass condensers properly located.

The filament circuit is laid out from the standpoint of obtaining the desired grid biasing potentials with the simplest possible filament circuit.

Detector and Audio Frequency Amplifier Circuits

The second detector is of the grid leak and grid condenser type because it has been found, by careful measurement, that the efficiency of the grid leak type of detector when using W. E. 215-A vacuum tubes is considerably greater than that of the negative grid bias type of detector up to inputs much greater than are likely to be obtained in practice. The disadvantage of the grid leak type of detector is that the output level obtainable is considerably lower than that from the negative grid bias type of detector. When one step of audio frequency amplification is provided, however, the output level obtainable from the receiver is ample for headphone operation. The detector and the audio frequency amplifier form a couplet which operate very well together, as the relative output levels are such that overloading occurs at about the same point in both tubes.

A by-pass condenser of the order of 0.001 microfarad is provided in the detector plate circuit in order to keep the output circuit of low impedance to the carrier frequency. In addition to obtaining an increase in detector efficiency, the condenser by-passes radio and intermediate frequencies which may be amplified by the audio frequency circuits and help to cause overloading of succeeding tubes. From the standpoint of detector efficiency an even larger capacity would have been advantageous, but it would have resulted in too great an attenuation of the higher audio frequencies.

Intermediate Frequency Oscillator

The intermediate frequency oscillator is of the tuned grid inductively coupled type and is so arranged that it may be turned off by means of a key in the plate supply. High efficiency in this oscillator was not desired nor was it necessary because of the fact that it operates at the frequency of the amplifier. The problem was not to obtain sufficient output from the oscillator, but to reduce the coupling from this oscillator to the detector to such a value that the detector would not be seriously overloaded. The by-pass condenser connected as shown reduced the input to the second detector from the intermediate frequency oscillator to about 0.5 volt at the grid of the second detector, which is approximately the value for maximum signal strength.

The capacity of the variable condenser was made only a small portion of the total tuning capacity as it was desired to have a frequency adjustment of only four or five thousand cycles.

It has been found that having this oscillator adjustable over a range of this order of magnitude is of value in differentiating between signals from two stations very close together in carrier frequency, as the radio frequency adjustments of the receiver may be set for the optimum strength of the desired station and the intermediate frequency oscillator adjusted so that the beat notes of the desired and undesired station may be most advantageously adjusted.

The coupling between the intermediate frequency and radio frequency oscillators is reduced to the lowest possible degree in order that the harmonics of the intermediate frequency oscillator will not beat with the fundamental of the radio frequency oscillator when it is adjusted over its operating range. With the intermediate frequency oscillator turned on, beat notes will occur for only two settings of the high frequency oscillator condenser corresponding to the frequencies 50 kc. above and below the carrier frequency.

Intermediate Frequency Amplifier

Prior to the development of the receiver being described, a very satisfactory intermediate frequency transformer had been developed. The frequency characteristic of one of these transformers is shown in Figure 11. This characteristic covers only the operating range, but it is also very important that audio frequencies and radio frequencies be not transmitted by the intermediate frequency transformer. It was found that when four of these transformers were used in an intermediate frequency amplifier, the characteristic obtained was quite different from the fourth power of the characteristic of a single transformer. This is to be expected, because the input impedance of a vacuum tube is a function not only of the grid to filament capacity but it is also affected by the make-up of its plate circuit. The input impedance of a grid leak detector tube is also very different from the input impedance of an amplifier tube.

A satisfactory over-all characteristic was obtained by balancing out some of the interstage coupling capacity. The balancing capacity is not only used to stabilize the amplifier and to reduce any tendency toward internal oscillation, but its proper adjustment determines the shape of the amplifier characteristic. The amplification of the receiver is controlled by means of a

potentiometer. The total resistance of this potentiometer is closely related to the value of the balancing capacity. The proper combination of these two values results in the desired characteristic. The potentiometric gain control shown is adjustable in ten steps having a voltage amplification ratio between them of approximately 2.5 to 1. The over-all amplification of the receiver for various steps of this amplification control is shown in Figure 12. The selectivity is greatest when the maximum amplification is used. This is a very desirable characteristic, as when a signal is so weak as to require the maximum amplification of the receiver a high degree of selectivity is desirable. The selectivity of the receiver is intentionally made considerably less than might be obtained in order to be able to receive signals when the carrier frequency changes slightly or is not absolutely accurately set by all transmitters.

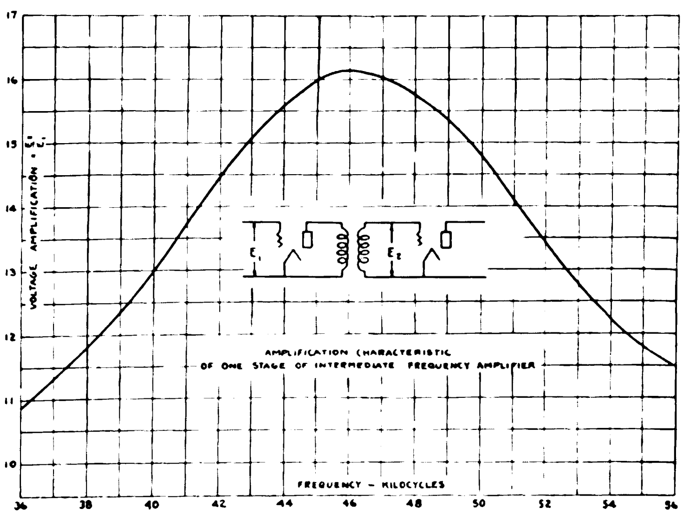

FIGURE 11

MEASUREMENT OF THE INTERMEDIATE FREQUENCY AMPLIFIER

For the measurement of the over-all intermediate frequency amplification, the circuit shown in Figure 13 is used. The input resistance R_1 assumes different values in accordance with the intermediate frequency amplification to be measured. The current thru the input resistance is kept constant at 1 milliampere. For measuring the amplification obtained on the two upper steps of the amplification control, the input resistance R_1 consists of a short straight piece of high resistance wire having

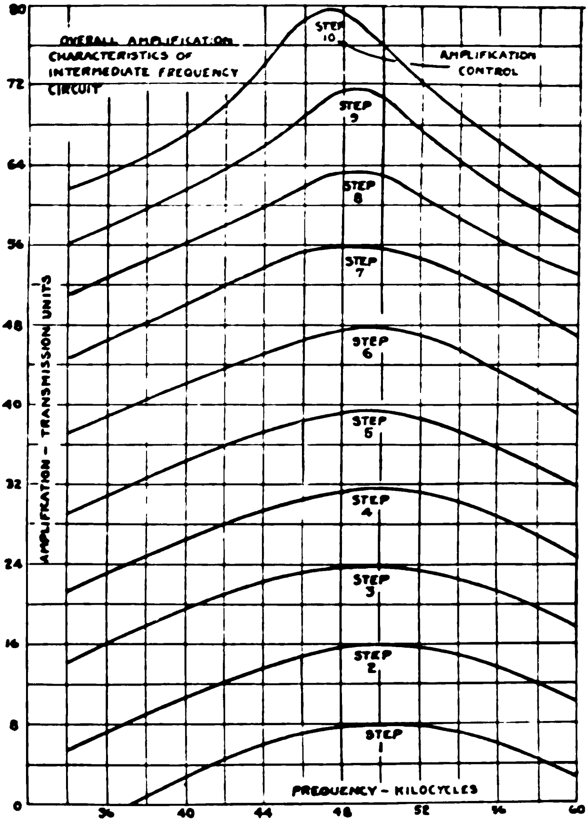

FIGURE 12

a resistance of 0.1 ohm mounted directly in the base of the short circuiting switch, this being the only practical way of eliminating undesirable pick-up. With a voltage amplification of 10,000, this means that 1 milliampere flowing thru this resistance will give an input of 1 volt to the grid of the second detector. When the amplification is reduced to 2,000 or less, resistance boxes may be used for the input resistance provided that precautions are observed to make all of the leads as short and direct as possible. It is essential that the input voltage to the receiver consist only of the drop across a definitely known resistance and that the current measured by the thermocouple should be the entire current thru this resistance and no other. The second detector, including its condenser and grid leak, is calibrated by connecting it directly across resistance R_1, which for this purpose consists of a variable resistance box having negligible inductance

and capacity at 50 kc. A complete calibration curve of the detector up to a voltage input corresponding to a change in the plate current of 200 microamperes is usually made. When measuring the amplification of the receiver, the current thru the input resistance is adjusted to the same value of 1 milliampere and the change in the detector plate current noted. This is most conveniently done by the use of a differential meter in which the normal space current of the detector is neutralized by a current from a separate battery flowing thru the proper resistance (see Figure 13). As the amplification is decreased by using the lower steps of the amplification control, the input resistance R_1 is increased so that the input voltage to the grid of the second detector is of the order of 0.5 to 1.0 volt. In all of these measurements it is very essential that the oscillator be thoroly shielded from the receiver so that with the switch across the resistance R_1 closed, no change in the plate current of the second detector may be noted when the oscillator is turned on and off, even tho the maximum amplification of the receiver is used. This method was used in measuring the frequency characteristic of the receiver.

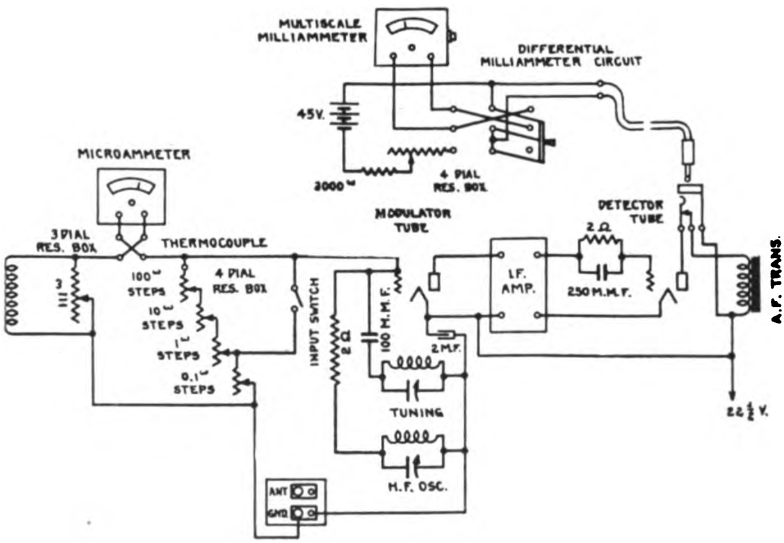

FIGURE 13

CONCLUSION

The transmitter and receiver meet the requirements and show what may be expected if the design is based on the results of intelligent co-ordination of theoretical and laboratory work. The

practical operation of the system shows that the range over water is considerably in excess of requirements, which seems to indicate that the shorter wave lengths will be very satisfactory for this class of service.

The authors wish to express their thanks to Messrs. P. H. Betts, A. W. Saunders, and A. E. Lanz for their valuable assistance in the preparation of this paper.

SUMMARY: This paper describes the transmitter and receiver recently developed for use by the United States Coast Guard. This apparatus is for operation on wave lengths between 100 and 200 meters. In describing the development of the transmitter a short summary of the various circuit considerations is included. The actual transmitter finally developed is also described together with its operating characteristics.

In considering the radio receiver the various problems to be met in the design of a radio receiver of this character are dealt with at some length. The frequency characteristics of the radio receiver, as developed, are shown, and the method of determining them is described in detail.

The transmitter and receiver performed very satisfactorily under conditions considerably more severe than will be met in actual service.

DESIGN OF TELEPHONE RECEIVERS FOR LOUD SPEAKING PURPOSES*

By

C. R. Hanna

(Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania)

I. Introduction

The development of telephone receivers to handle without serious distortion the relatively large amounts of power required for loud speaking purposes has presented many difficulties. The simple receiver of bi-polar construction sufficed as long as it had to take care of the small amount of power used in ordinary telephone practice, but considerable modification, if not a complete change of design, became necessary when loud speakers came into existence.

At the present time three types of magnetic driving mechanisms are employed for loud speakers: (1) The moving iron or electromagnetic type; (2) The moving coil or electrodynamic type, including all those forms in which conductors carrying primary currents are displaced in a steady magnetic field; and (3) The induction or eddy current type, including all those in which conductors carrying secondary or induced currents are displaced in a steady field.

The present paper will deal only with the moving iron or the electromagnetic type of driving element. This is by far the most common. It is also the least expensive, probably because its steady flux is readily supplied by permanent magnets instead of electromagnets. It depends for its operation upon the variation in pull of an electromagnet or system of electromagnets upon an iron armature or diaphragm. Several types of electromagnetic constructions will be discussed, and then a detailed consideration of the design of a new type recently developed will be given.

Simple Bi-polar Receiver

The familiar two-pole receiver is shown in Figure 1. Many

* Received by the Editor, January 23, 1925. Presented before The Institute of Radio Engineers, New York, February 4, 1925.

papers and books have been written describing its operation and characteristics. The reader is referred to A. E. Kennelly's "Electrical Vibration Instruments," which covers the subject very thoroly and which gives a complete bibliography. A few relations showing first order effects will here be given to show the limitations of this type of element for loud-speaking telephones.

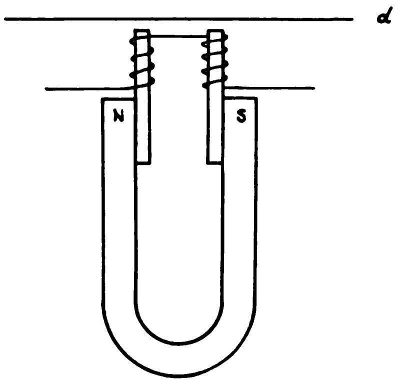

FIGURE 1—Simple Receiver

If Φ is the steady component of flux crossing the air gaps and $\phi \sin \omega t$ represents the superposed alternating component of flux due to a sine wave of current in the windings, the pull on the diaphragm will be proportional to

$$(\Phi+\phi \sin \omega t)^2 = \Phi^2+2\,\Phi\,\phi \sin \omega t+\phi^2 \sin^2 \omega t$$
$$= \Phi^2+2\,\Phi\,\phi \sin \omega t+\phi^2\left(\frac{1-\cos 2\,\omega t}{2}\right)$$
$$= \Phi^2+\frac{\phi^2}{2}+2\Phi\,\phi \sin \omega t-\frac{\phi^2}{2}\cos 2\,\omega t$$

It is seen that the pull is made of three components: (1) a steady component $\left(\Phi+\frac{\phi^2}{2}\right)$; (2) a single frequency component $(2\,\Phi\,\phi \sin \omega t)$ and (3) a double frequency component $\frac{\phi^2}{2}\cos 2\,\omega t$.

The steady component produces no undesirable effect, but the double frequency pull gives rise to an extraneous overtone which is quite objectionable. It is seen that if ϕ is small compared to Φ the double frequency term will be negligible compared to the single frequency term. In loud speakers, however, ϕ has to be large in order to produce sufficient variation of force, and this

in turn requires that Φ shall be much larger, in order to make the single frequency component of force predominate.

In the simple receiver the limit to the amount of steady flux that can be employed is determined by the saturation of the diaphragm. Increasing the thickness of the diaphragm allows the use of more steady flux and, therefore, makes the device capable of handling more power without the double frequency distortion referred to above. This procedure is objectionable, however, because it increases the mass and rigidity of the diaphragm, both of which are undesirable because of the non-uniformity of response they cause. Increase in power is, therefore, incompatible with good quality of response in the simple receiver.

Another source of distortion inherent to the simple receiver when used for loud-speaking purposes is the unsymmetrical displacement of the diaphragm away from and toward the poles with respect to its mean position. This is shown in Figure 2 as a static characteristic of displacement against current in the coils. For the large displacements necessary in loud-speaking receivers, this non-linear distortion is very objectionable because of the extraneous overtones it causes.

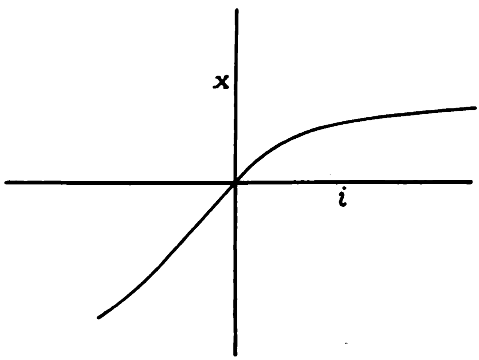

FIGURE 2—Static Characteristic of Simple Receiver

It may be said, however, that bi-polar receiver constructions are being employed with fair success for loud speakers when only a moderate amount of output is required. Fairly large diaphragms are employed so that the displacements do not have to be as great for a given amount of power as with smaller diaphragms.

BALANCED ARMATURE UNIT

To overcome the disadvantages inherent to the simple re-

ceivers, the four-pole balanced armature receiver was developed. Two constructions are shown in Figure 3, the operation of the two being the same. When current passes thru the coil (or coils) poles 1 and 4 are strengthened, while 2 and 3 are weakened (or *vice versa*), causing the armature to be pulled in the direction of the stronger pair of poles. When in its normal position, the armature carries little or no steady flux. The steady flux may, therefore, be large without saturation of the armature. It is not, however, to be inferred that with this construction there is no limit to the steady flux but saturation of the poles. If the field is too strong, the armature will not stay in its mid-position. This can be shown as follows: Suppose the armature is rotated slightly clockwise. Poles 1 and 4 will now have a greater pull on the armature than poles 2 and 3 because of the reduced air gap. If the difference between these two forces is not less than the restoring force called into play by the deflection of the moving system, the armature will be pulled over to poles 1 and 4. The armature is continually being deflected from its mid position, and so the strength of the poles can not exceed a certain value without causing instability. This limit, however, is much higher than in the simple receiver where diaphragm saturation is involved.

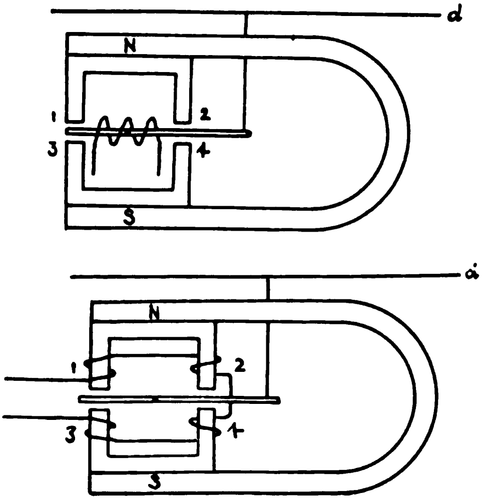

FIGURE 3—Balanced Armature Receivers

The chief advantage of a strong field in this type of receiver is that the sensitivity is greater. The double frequency component of force, so objectionable in the simple receiver, is not

present in this type if the armature is closely balanced. That there is no double frequency pull may be shown as follows: The pull caused by poles 1 and 4 is proportional to

$$(\Phi + \phi \, sin \, \omega t)^2$$

and the pull of poles 2 and 3 is proportional to

$$(\Phi - \phi \, sin \, \omega t)^2$$

where Φ represents the steady component of flux and $\phi \, sin \, \omega t$ the superposed alternating flux. The resultant rotating force (F) is given by the difference between the above expressions.

$$(\Phi + \phi \, sin \, \omega t)^2 - (\Phi - \phi \, sin \, \omega t)^2$$
$$= (\Phi^2 + 2 \, \Phi\phi \, sin \, \omega t + \phi^2 sin^2 \, \omega t) - (\Phi^2 - 2 \, \Phi \, \phi \, sin \, \omega t + \phi^2 sin^2 \, \omega t)$$
$$= 4 \, \Phi \, \phi \, sin \, \omega t$$

It is seen that the steady component and the double frequency component of pull both disappear, leaving only the single frequency.

Because of the four-pole arrangement, the deflections of the armature are symmetrical with respect to its mean position. That is, the static characteristic of displacement against current is as in Figure 4, where the range of linearity is roughly double that of the simple receiver having the same air gap clearance.

The fact that the balanced armature receiver will stand larger variations of flux without double frequency distortion, and larger diaphragm displacements without the other type of non-linear distortion mentioned, and the fact that it allows greater steady fields without saturation of the moving part, makes it much better suited for loud-speaker work than the simple receiver.

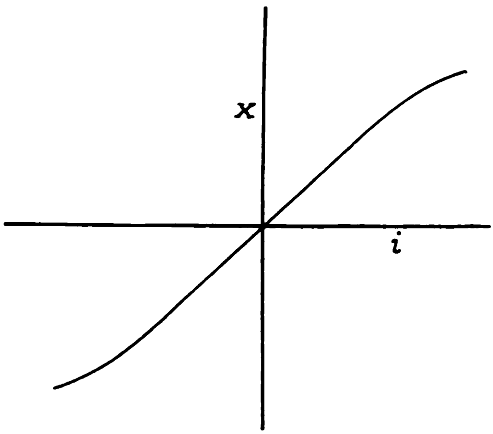

FIGURE 4—Static Characteristic of Balanced Receiver

The chief disadvantage of the balanced armature unit is its complicated vibrating system. The diaphragm has several resonances, and also the armature has one or more in the audible range unless made very heavy. The result is that the two coupled mechanical systems give rise to a rather bad anti-resonance, which is difficult to smooth out. If the balanced principle could be employed without the use of an armature, not only this disadvantage would be overcome, but a receiver with a lighter vibrating system would be obtained. Such a receiver has been developed and has been termed the balanced diaphragm type. The body of the paper will be concerned with the details in the design of this new loud-speaking receiver.

II. Balanced Diaphragm Receiver

The construction of the balanced diaphragm receiver is shown in Figures 5 and 6. Two pairs of poles are disposed on opposite sides of the disphragm in such a way that the steady flux passes across the air gaps without going lengthwise thru the diaphragm. Essentially the construction consists of two simple bi-polar receivers actuating a common diaphragm. Normally both pairs

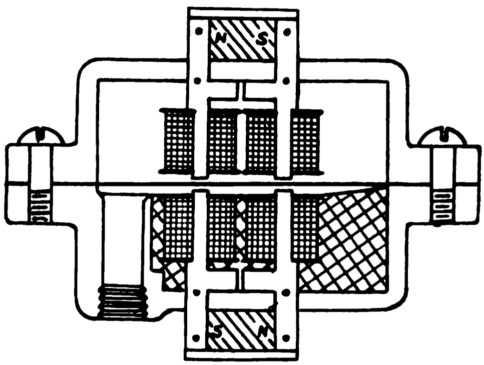

Figure 5—Balanced Diaphragm Receiver

of poles pull equally on the diaphragm. The windings are so connected that current will strengthen one pair (say 1 and 2) and weaken the other, or *vice versa*, causing a force on the diaphragm in the direction of the stronger pair of poles. In the construction shown, two short blocks of high coercive force magnet steel are used for supplying the steady flux. These are placed outside the case for ease of magnetizing. The bridge or shunting tath below the coils on each half of the receiver is of low reluc-

tance compared to the main gaps, and allows the varying flux component an easy path around the high reluctance permanent magnet. This leakage path is, of course, a shunt to the steady

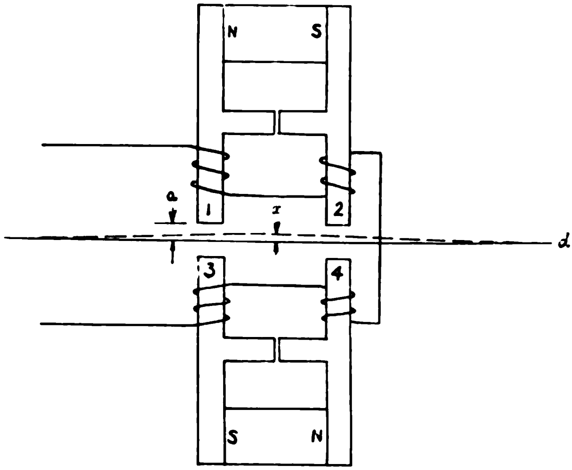

FIGURE 6—Balanced Diaphragm Receiver

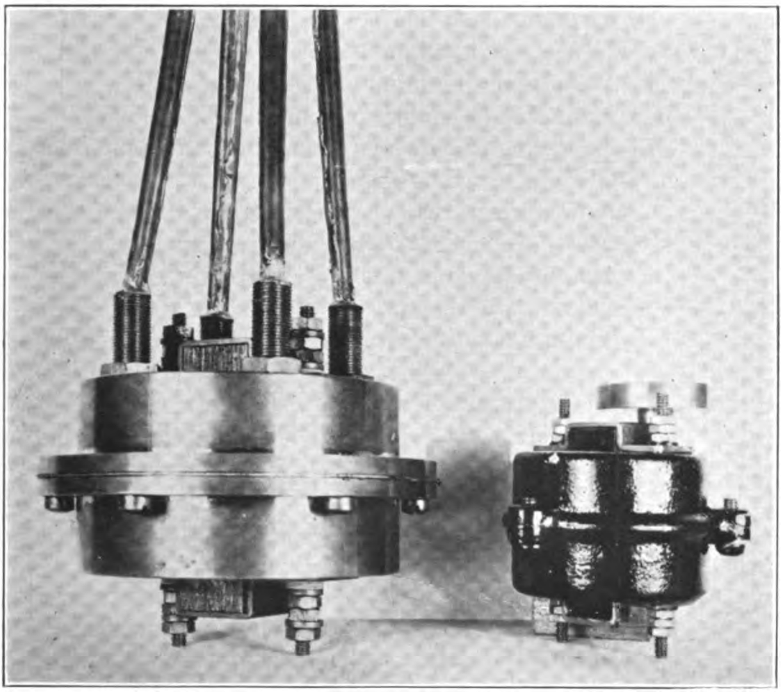

FIGURE 7

flux and requires that the magnet section be large enough to supply both the leakage flux and the useful flux. The poles are built up of T-shape laminations riveted together at the lower ends where there is little or no variation of flux. Because of the pole construction the sound outlet is off center as shown. Wax is used to fill up the one-half of the receiver so that the air chamber next to the diaphragm and ahead of the orifice is small. This space is usually made approximately conical in shape with an altitude of from ten to twenty mils. The advantages of the balanced diaphragm construction are the same as those given in the discussion of the balanced armature with the added feature that no armature is employed to obtain those advantages.

Figure 7 shows two sizes of the balanced diaphragm loud-speaking receiver, one for use in the home and one for power work. Design information has been worked out whereby different sizes of receivers having the greatest possible sensitivity may be obtained without resorting to the cut-and-dry method. The various constants for the receiver will now be determined and the details in designing a receiver given.

The following notation will be used:

A = Pole area (cm.^{2})
a = Air gap length (cm.)
x = Deflection diaphragm at center (cm.)
$\dot{x}$ = Velocity of diaphragm at center (cm./sec.)
M_o = Mmf. due to one permanent magnet (gilberts)
M = Mmf. due to current i in one pair of coils (gilberts)
i = Current in windings (amperes)
n = Number of turns per pair of coils
B = Flux density in air gap (gausses)
A_d = Diaphragm area (cm.^{2})
t = Diaphragm thickness (cm.)
e = Young's modulus for diaphragm material (dynes/cm.)
δ = Density of diaphragm material (gm./cm.^{3})
f_o = Resonance frequency of diaphragm if not in the presence of magnetic field (cycles/sec.)
f_r = Resonance frequency in presence of magnetic field
m = Equivalent mass of vibrating diaphragm, referred to acceleration at center (gm.)
r = Damping, referred to velocity at center of diaphragm, dynes/(cm./sec.)
S = Stiffness of diaphragm not in presence of magnetic field, referred to deflection at center (dynes/cm.)

S_m = Magnetic reduction of stiffness (dynes/cm.)
L = Inductance of receiver (henrys)
K = Force factor or force per unit current in windings (dynes/ampere)

Referring to Figure 6, suppose the current at any instant is flowing so as to increase the strength of poles 1 and 2 and decrease 3 and 4, and that the diaphragm is displaced a distance x which is practically the same at the poles as at the center. Then, neglecting all reluctances excepr those of the main air gaps

$$B_{12} = \frac{M_o + M}{2(a-x)}$$

$$B_{34} = \frac{M_o - M}{2(a+x)}$$

The resultant force on the diaphragm

$$F = \frac{1}{8\pi} B_{12}^2 \,.\, 2A - \frac{1}{8\pi} B_{34}^2 \cdot 2A =$$

$$= \frac{A}{16\pi}\left[\left(\frac{M_o+M}{a-x}\right)^2 - \left(\frac{M_o-M}{a+x}\right)^2\right]$$

$$= \frac{A}{16\pi}\left[\frac{M_o+M}{a-x} - \frac{M_o-M}{a+x}\right]\left[\frac{M_o+M}{a-x} + \frac{M_o-M}{a+x}\right]$$

$$= \frac{A}{16\pi}\left[\frac{2(a\,M + x\,M_o)}{a^2-x^2}\right]\left[\frac{2(a\,M_o + x\,M)}{a^2-x^2}\right]$$

If x is small compared with a, and M small compared with M_o

$$F = \frac{A}{4\pi} \frac{(a\,M + x\,M_o)\,(a\,M_o)}{a^4}$$

$$= \frac{A\,M_o}{4\pi a^2} M + \frac{A\,M_o^2}{4\pi a^3} x\,. \qquad (1)$$

The force acting on the diaphragm is made up to two components, one proportional to and in phase with the mmf. due to current in the coils, and one proportional to and in phase with the displacement of the diaphragm. This, of course, neglects the effect of eddy currents and hysteresis in the poles and diaphragm. The first term determines the force per unit current or force factor

$$K = \frac{F_i}{i} = \frac{A\,M_o}{4\pi a^2} \frac{M}{i}$$

$$= \frac{A\,M_o}{4\pi a^2} \times 0.4\,\pi\,n$$

$$= \frac{A\,M_o\,n}{10\,a^2} \text{ dynes per ampere.} \qquad (2)$$

The force proportional to x is in effect a reduction of diaphragm rigidity since it is opposed to the restoring force. This reduction of stiffness is

$$S_m = \frac{F_x}{x} = \frac{A\,M_o^2}{4\,\pi\,a^3}\ \text{dynes/cm.} \tag{3}$$

These two quantities are important, and will be referred to constantly thruout the paper.

The problem in design is, of course, to secure the largest value of K for a receiver having a specified impedance. From equation (2) it is seen that increasing the steady magnetomotive force M_o will accomplish this, but from equation (3) we see that there is a limit to M_o because of the instability of the diaphragm when S_m is greater than the diaphragm rigidity. Likewise, to increase A or reduce a will increase K, but there is a limit here because of the corresponding increase of S_m. Hence the upper limit of S_m is a quantity which partially determines the greatest value of K.

It will now be shown that for operation in a given tube circuit the inductive reactance of the receiver should equal the tube impedance at the frequency where the power is to be optimum. It is not meant by this that the frequency response curve is to show a maximum at this frequency, but that conditions are to be so chosen that the greatest response capable of being obtained at this frequency results. In all receivers of the electromagnetic type the inductive reactance predominates at the higher frequencies (where the optimum is usually desired), and so this will be considered as the whole impedance. If $\mu\,e_g$ is the effective voltage acting in the plate circuit, R_p being the tube impedance

$$i = \frac{\mu\,e_g}{\sqrt{R_p^2 + \omega^2 L^2}}$$

It was seen in the first part of the paper that the pull on the diaphragm is proportional to the first power of the variable component of flux in the air gaps. This is given by

$$F \propto \phi \propto \frac{L\,i}{N} \qquad N = \text{total turns in series}$$

But

$$L \propto N^2$$

$$F \propto \sqrt{L}\,i$$

Substituting the above value of i

$$F \propto \mu\,e_g \sqrt{\frac{L}{R_p^2 + \omega^2 L^2}}$$

This is maximum if

$$\omega L = R_p \tag{4}$$

We have then the two determining conditions: (1) for a given diaphragm the magnetic reduction of rigidity must be less than the diaphragm rigidity; and (2) for operation in a given tube circuit the inductance is fixed by the relation that the inductive reactance should equal the tube impedance at the frequency where the optimum is to be obtained. It will now be shown that if, in the design of a receiver, these two quantities are obtained in the values prescribed, the force factor K will be independent of the pole area or air gap length. This neglects the effects of iron saturation. We have

$$S_m = \frac{A\,M_o^2}{4\,\pi\,a^3} \tag{5}$$

$$L = 2\left(\frac{0.4\,\pi\,n^2\,A}{2\,a}\right)10^{-8} \tag{6}$$

$$K = \frac{A\,M_o\,n}{10\,a^2} \tag{7}$$

From (5) $$A\,M_o = \sqrt{4\,\pi\,a^3\,A\,S_m}$$

From (6) $$n = \sqrt{\frac{10^8\,L a}{0.4\,\pi\,A}}$$

Substituting in (7)

$$K = \frac{\sqrt{4\,\pi\,a^3\,A\,S_m}}{10\,a^2}\sqrt{\frac{10^8\,L\,a}{0.4\,\pi A}}$$

$$= \sqrt{10^7\,L\,S_m} \tag{8}$$

This conclusion may be seen also in a qualitative way, for if the air gap is increased or the pole area is reduced, the turns must be increased to give the same inductance and the permanent mmf. must be increased to give the same magnetic reduction of rigidity. Both of these increases compensate for the reduction of K due to the greater gap or smaller pole area.

The importance of the foregoing conclusion will be evident in the design of receivers of this type. The procedure is simply as follows:

1. Select a diaphragm large enough to radiate the amount of power required and of the correct thickness to give the desired resonance frequency. This latter should be calculated higher than the frequency desired because of the reduction of rigidity due to the magnetic field, which usually may be 50 percent of the intrinsic diaphragm stiffness.

2. Fix the air gap at a convenient value, say at least four times the maximum displacement of the diaphragm.

3. Choose a convenient pole area, taking into account only the requirement that the poles shall not be saturated to the point where their reluctance is comparable to that of the air gaps.

4. Use the strongest permanent magnet that will still allow the diaphragm to remain in its mid position with stability.

5. Calculate the number of turns to give the required inductance. Equation (8) predicts that variations in the dimensions, which were chosen more or less arbitrarily, will not affect the sensitivity of the resulting receiver. Thus the design of such receivers is made quite simple, and with a degree of assurance that the greatest possible sensitivity will be obtained.

III. Diaphragms and Horns

Before dealing with an actual design, some of the properties of diaphragms and horns should be reviewed. It can be shown that if a loud speaker is to operate between the limiting frequencies f_1 and f_2, its resonance frequency should be roughly

$$f_r = \sqrt{f_1 f_2} \tag{9}$$

Rayleigh gives the following approximate formula for the fundamental resonance frequency of a circular diaphragm clamped around its circumference:

$$f_o = 1.48 \frac{t}{A_d} \sqrt{\frac{e}{\delta}} \tag{10}$$

where $\sqrt{\frac{e}{\delta}}$ is the velocity of sound in the metal, usually about 5.10^5 cm./sec. Hence

$$f_o = 7.4 \cdot 10^5 \frac{t}{A_d} \tag{11}$$

It can be shown that the equivalent mass of vibration of a clamped diaphragm for its fundamental mode is roughly one-fourth its total mass. From these last two relations, then, the rigidity may be calculated, using

$$f_o = \frac{1}{2\pi} \sqrt{\frac{S}{m}} \tag{12}$$

If the reduction of rigidity due to the presence of the steady magnetic field is of the order of 50 percent of the intrinsic diaphragm stiffness, the actual resonance frequency of the system will be

$$f_r = \frac{\sqrt{2}}{2} f_o$$

From equation (10) it is seen that the ratio of diaphragm thickness to area must be constant for a receiver having a given resonance frequency. The mass of the diaphragm (and hence its equivalent mass) will, therefore, vary as the square of the area. From equation (12) the rigidity will also vary as the square of the area because of the constant ratio of S to m. The magnetic reduction of rigidity may, therefore, be greater for large diaphragms. Referring to equation (8) for the force factor of a receiver, it might be thought that since K is larger when large diaphragms are used, the over-all sensitivity of large loud speakers would be greater. Such would be the case if a horn providing the same acoustic load on the diaphragm were employed, but with heavier and stiffer diaphragms it is necessary that this loading due to the horn be increased so as to give the same degree of uniformity in response over the frequency range. The loading or radiation damping required for a given degree of uniformity is proportional to the mass or to the stiffness of the diaphragm.[1]

If, as indicated in the paper referred to, the radiation damping r is large compared to other losses in the diaphragm, the power radiated at the resonance frequency will be

$$W \propto \frac{K^2}{r}$$

But $$K \propto \sqrt{S_m} \propto A_d$$

and $$r \propto m \propto A_d{}^2$$

Hence W is constant. Thus it may be reasonably expected that all sizes of the balanced diaphragm receiver, if fitted with properly designed horns so as to give the same uniformity of response, will have the same over-all sensitivity.

In the paper referred to, the following approximate formula for radiation damping or loading is given

$$r = 4.6\frac{A_d{}^2}{A_o} \tag{13}$$

where A_o is the initial area of the horn. It is seen that if A_o is held constant, the radiation damping will vary with the diaphragm area in just the right manner to give the same uniformity of response independent of diaphragm size.

The power that a diaphragm radiates using a properly designed horn is proportional to the square of its area and to the square of its velocity. If it is assumed that the maximum allowable deflection of a diaphragm is proportional to its diameter

[1] See paper "The Function and Design of Horns for Loud Speakers," by C. R. Hanna and J. Slepian, "Journal A. I. E. E.," March, 1924.

(hence to the square root of its area), then since the velocity at a given frequency is proportional to the deflection, the greatest power output from a given diaphragm will be proportional to the cube of the area. That is,

$$W \propto \dot{x}^2 \cdot A_d^2$$
$$\dot{x} \propto x \propto \sqrt{A_d}$$
$$W \propto A_d^3$$

This is significant in the design of power loud speakers, for if the diameter of a diaphragm is doubled, the amount of power it can be made to radiate without serious distortion will be increased 64 times. The distance range varying as the square root of the power will be increased eight times. The large size unit shown in the photograph has a 3-inch diaphragm, and is capable of radiating about a watt of sound in speech and music. Altho this may seem small, a watt of sound is probably all that a fifty-piece band can produce. It is almost incredible that a 3-inch diaphragm moving not more than four or five thousandths of an inch will produce as much sound as a good-sized band, but such is the case. This is accounted for by the large radiation resistance or damping imposed on the diaphragm by a properly designed horn. Such sounds have been heard with great loudness about a half mile away and because of the logarithmic character of the ear's impression of intensity, could probably have been heard distinctly over two miles distance if there were no obstructions in the way.

For several reasons it has not been found desirable to employ diaphragms larger than 3 inches in diameter. First, the volume of the air chamber immediately above the diaphragm cannot be kept sufficiently small without making the height of the space less than a practicable value. (In the paper referred to it was shown that unless this volume is small, the radiation at the higher frequencies is materially reduced, because of the fact that the air is compressed in the space instead of being forced into the horn.) Second, because of the greater length of path across the surface of the diaphragm to the outlet, space resonances occur at frequencies within the working range unless several outlets are distributed over the diaphragm surface in such a way as to keep the effective length of path small. With large diaphragms the number of outlets required necessitates that each one shall be excessively small if the total outlet area is prescribed according to information given in the paper on horn design referred to above. There are four outlets in the large receiver shown in Figure 7, each one being 3/16 inch in diameter.

Diaphragms 1½ to 2 inches in diameter have been found large enough to radiate all the sound necessary for home use. The smaller of the two receivers shown in the photograph has a 1¾-inch diaphragm.

IV. Details of Receiver Design

Suppose a power loud speaker with the following characteristics is required:

1. Maximum power output of 1 watt at 500 cycles.
2. Frequency range 50–4,000 cycles.
3. Impedance of proper value for 5,000-ohm tube.

If a 3-inch diaphragm ($A_d = 45.6$ cm.²) is employed and the total orifice area is equivalent to a ⅜-inch diameter hole, the radiation damping imposed on the diaphragm by a properly designed exponential horn will be

$$r = 4.6\frac{A_d^2}{A_o} = 13{,}500 \text{ dynes}/(\text{cm.}/\text{sec.})$$

The maximum deflection at 500 cycles in order to radiate 1 watt or 10^7 ergs per second is obtained as follows:

$$\frac{1}{2}\dot{x}^2 r = \frac{1}{2}(\omega x)^2 r = W$$

$$\frac{1}{2}(2\pi \cdot 500 \cdot x)^2\, 13{,}500 = 10^7$$

$$x = 0.0123 \text{ cm.} = 0.0048 \text{ inch}$$

The air gap should be approximately four times this or 0.020 inch. The resonance frequency of the diaphragm should be

$$f_r = \sqrt{50 \times 4000} = 450 \text{ cycles.}$$

If the magnetic reduction of rigidity is half the intrinsic diaphragm stiffness, f_r will be $\frac{\sqrt{2}}{2} \cdot f_o$, so that the resonant frequency of the diaphragm when not in the magnetic field should be

$$f_o = 450\sqrt{2} = 635 \text{ cycles.}$$

The thickness of the diaphragm is obtained from equation (11)

$$635 = 7.4 \cdot 10^5 \frac{t}{45.6}$$

$$t = 0.0392 \text{ cm.} = 0.0154 \text{ inch, or say } 0.015 \text{ inch.}$$

The actual value of f_o for $t = 015$ inch is 620 cycles/sec.

The equivalent mass of the diaphragm for the fundamental mode of vibration taking $\delta = 7.8$ is

$$m=\frac{1}{4}(45.6\times 2.54\times 0.015)7.8=3.4 \text{ gm.}$$

From equation (12) the stiffness may be calculated

$$S=(2\pi f_o)^2\, m=(2\pi\ .\ 620)^2\times 3.4$$
$$=51.3\times 10^6 \text{ dynes/cm.}$$

The magnet and pole construction should be designed so as to give a magnetic reduction of stiffness about half the above, of

$$S_m=25\times 10^6 \text{ (say)}$$

The actual resonance frequency in the field will be

$$f_r=\sqrt{\frac{51.3-25}{51.3}}\times 620=445 \text{ cycles/sec.}$$

The pole area is the only other arbitrarily chosen value. Let each pole be 3/16 inch$\times$⅝ inch and later we shall determrine whether the steady flux required to give the above value of S_m saturates the poles or not.

$$A=3/16\times 5/8\times 6.45=0.756 \text{ cm.}^2$$
$$a=0.020\times 2.54=0.0508 \text{ cm.}$$

From equation (3)

$$M_o=\sqrt{\frac{4\pi a^3 S_m}{A}}=\sqrt{\frac{4\pi(0.0508)^3\, 25.10^4}{0.756}}$$
$$=233 \text{ gilberts.}$$

The density of the steady flux in the poles will be

$$B_o=\frac{M_o}{2a}=\frac{233}{2\times 0.0508}=2{,}300 \text{ gausses}$$

which is sufficiently low.

The bridge or leakage path should have a reluctance of about one-fourth that of the two main air gaps. A path 3/16 inch$\times$¾ inch in area and 0.0125 in length has this reluctance. The density of flux in the bridge gap will be

$$B_s=\frac{233}{0.0125\times 2.54}=7{,}350 \text{ gausses}$$

which is also sufficiently low. The total flux delivery from the magnet must be five times that in the main poles or

$$\Phi_m=5\, B_o\, A=5\ .\ 2{,}300\ .\ 0.756$$
$$=8{,}700 \text{ maxwells.}$$

Each magnet of the receiver must deliver 8,700 lines at 233 gilberts mmf. The problem of designing a permanent magnet

capable of delivering a certain flux at a gven magnetomotive force is not difficult if the hysteresis loop for the particular steel is had. With cobalt, tungsten, or chromium steels, the flux density in the steel for the most economical magnet should be from 5,000 to 6,000 gausses. At this density the mmf. per centimeter length of magnet is from 120 to 160 gilberts for cobalt steel, and 40 to 50 gilberts for tungsten or chromium steel. Suppose a sample of cobalt steel when worked at a density of 5,000 gausses supplies 150 gilberts mmf. per centimeter length. For the particular application where 8,700 maxwells at a magnetomotive force of 233 gilberts are required, the length of magnet

$$l_m = \frac{233}{150} = 1.55 \text{ cm.} = 0.61 \text{ inch, say } 5/8 \text{ inch}$$

and the area

$$A_m = \frac{8700}{5000} = 1.74 \text{ cm}^2.$$

A magnet ¾ inch × ¾ inch in section has 1.81 cm.² area, which is approximately correct. If tungsten or chromium steel is used, the length should be about 3½ times the above, and because of the greater leakage or fringing, the section should be greater, especially at the center of the magnet where the total flux is greater. Cobalt steel magnets for applications of this kind are usually about ¼ or 1/5 as large as magnets of other steels. The receiver construction is also simplified if the magnet is short enough to go between the poles as shown in Figure 5. This is a factor which usually determines the area of the poles of the receiver. For example, if, with the pole area chosen, the value of M_o had been greater than that which could be obtained with a convenient length of magnet, the area of pole could be increased and the required value of M_o thereby reduced.

The remaining problem is to determine the proper windings. If at 3,000 cycles the response is to be as great as possible for that frequency, the inductance at that frequency is obtained from

$$\omega L = R_p$$

$$L = \frac{5000}{2\pi \cdot 3000} = 0.266 \text{ henry.}$$

In most laminated pole receivers tested, it has been found that the inductance decreases with frequency, approaching a limiting value of about half the low frequency inductance. This limit is usually reached below 3,000 cycles, and so the low frequency inductance of the receiver should be about double the above or

$$L_o = 0.5 \text{ henry (approx.)}$$

The number of turns per pair of poles may be calculated from

$$L_o = 2\left(\frac{0.4\,\pi\,n^2\,A}{2\,a}\right)10^{-8}$$

$$0.5 = 2\left(\frac{0.4\,\pi\,n^2 \cdot 0.756}{2\times 0.0508}\right)10^{-8}$$

$$n = 1{,}640 \text{ turns.}$$

With reasonable winding space, the ohmic resistance of the receiver is always small compared to the tube impedance, and so its value does not concern us, excepting as the windings might heat with the relatively large amounts of power employed. The loud speaker is usually in the secondary circuit of a transformer for insulation from the high voltage used in power amplifiers, and hence carries only the voice currents. Because of the great fluctuations of intensity, the average power in speech and music is usually so low compared to the peak value as to cause little or no heating of the windings or cores. So a size of wire is chosen which will allow the proper number of turns in a convenient winding space, and the matter of resistance forgotten. In the receiver here designed, 820 turns of number 36 wire can be conveniently wound on each pole if the winding section is ⅜ inch long by about 3/16 inch thick.

The principal dimensions and characteristics of the receiver will be tabulated:

Diaphragm 0.015 inch thick by 3 inch diameter.

$f_o = 620$ cycles/sec.

$m = 3.4$ grams

$S = 51.3\times 10^6$ dynes/cm.

$S_m = 25\times 10^6$ dynes/cm.

$f_r = 445$ cycles/sec.

$a = 0.020$ inch

$A = 3/16$ inch × ⅝ inch

$M_o = 233$ gilberts

Leakage path 3/16 inch × ¾ inch by 0.0125 inch long

Total flux delivery from magnet, 8,700 maxwells

Magnets, cobalt steel, ⅜ inch × ¾ inch × ⅝ inch long.

Inductance 0.5 henry at low frequencies

$n = 1{,}640$ turns per pair poles

Number 36 wire in space ⅜ inch × 3/16 inch

The force per unit current in the windings is given by equation (8)

$$K = \sqrt{10^7\,L\,S_m} = \sqrt{10^7 \cdot 0.5\times 25 \cdot 10^6}$$

$$= 11.2\times 10^6 \text{ dynes/ampere.}$$

This latter is the direct current force factor and will be found to decrease to a limiting value of about half with increase in frequency for most laminated pole receivers. With solid poles the percentage decrease in force factor is greater and the rate of decrease is more rapid with increase in frequency. Laminations of silicon steel 0.014 inch thick have been used with good results.

Details in the mechanical design of the receiver will not be discussed except to say that the diaphragm must be rigidly clamped with no metal to metal contacts. Paper, cloth, or rubber washers are used on each side of the diaphragm to prevent rattling. In the large size receivers rubber dam 7 to 10 mils thick has proven the best for washers. The air gaps should of course be accurate within close limits and the diaphragm flat if the full benefit of the balanced construction is to be obtained.

V. Experimental Work

By means of motional impedance diagrams some of the predicted characteristics of receivers can be checked experimentally. The equations for force factor and reduction of stiffness are found to give results which agree closely with those obtained by experiment. These equations are

$$K_o = \sqrt{10^7 L_o S_m}$$

$$S_m = \frac{A M_o^2}{4\pi a^3}$$

A receiver having the following principal dimensions was used in the test:

Diaphragm — 0.007 inch thick × 1¾ inch diameter.
Poles 3/32 inch × 5/16 inch. $A = 0.189$ cm.^{2}
Air gap 0.0125 inch. $a = 0.0318$ cm.

The equivalent mass of the vibrating diaphragm, assuming it to be one-fourth its total mass, is $m = 0.54$ gm.

The magnetomotive force of the permanent magnet in each half of the receiver, as determined by measuring the pull on a thick iron armature at a known distance from the main poles was found to be

$$M_o = 107 \text{ gilberts.}$$

Using a vacuum tube oscillator and an alternating current bridge, the resistance and reactance of the receiver were measured at different frequencies and are plotted in Figure 8. The smooth asymptotic curves are the estimated resistance and reactance if the diaphragm were not permitted to move. The differences

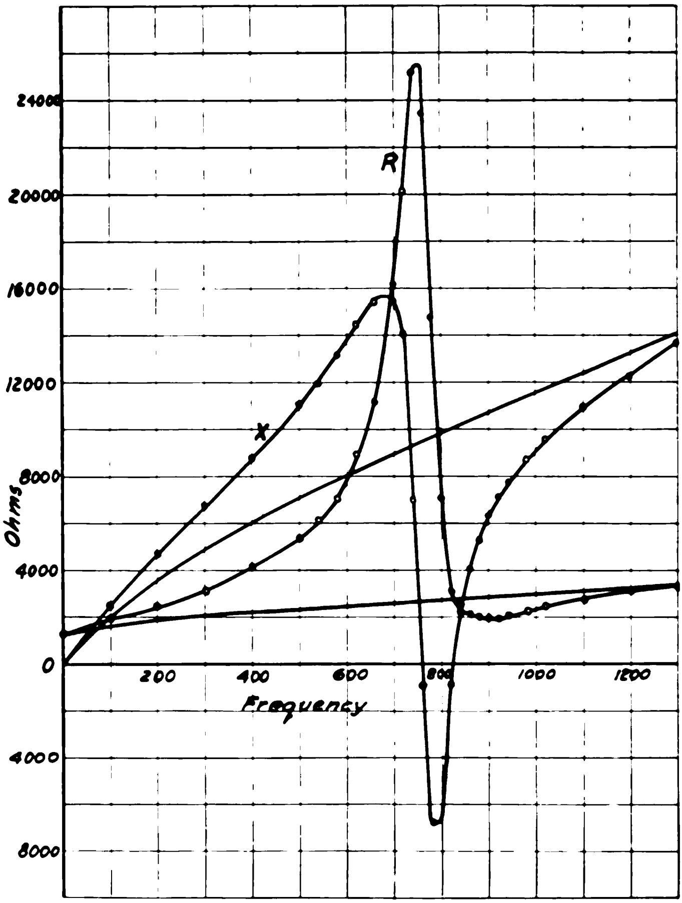

FIGURE 8—Resistance and Reactance Curves

between the values on the actual curves and the asymptotic curves are the motional resistance and the motional reactance. These two values may be plotted as a vector, the arrow of which will be found to describe a circle as the frequency is varied. This motional impedance circle for the receiver tested is shown in Figure 9. The resonance frequency of the diaphragm is 760 cycles, where the motional impedance vector is maximum.

The damping constant r of the receiver diaphragm is obtained from the logarithmic decrement

$$\Delta = \frac{r}{2\,m} = \pi(f_2 - f_1)$$

when f_1 and f_2 are frequencies corresponding to the extremities

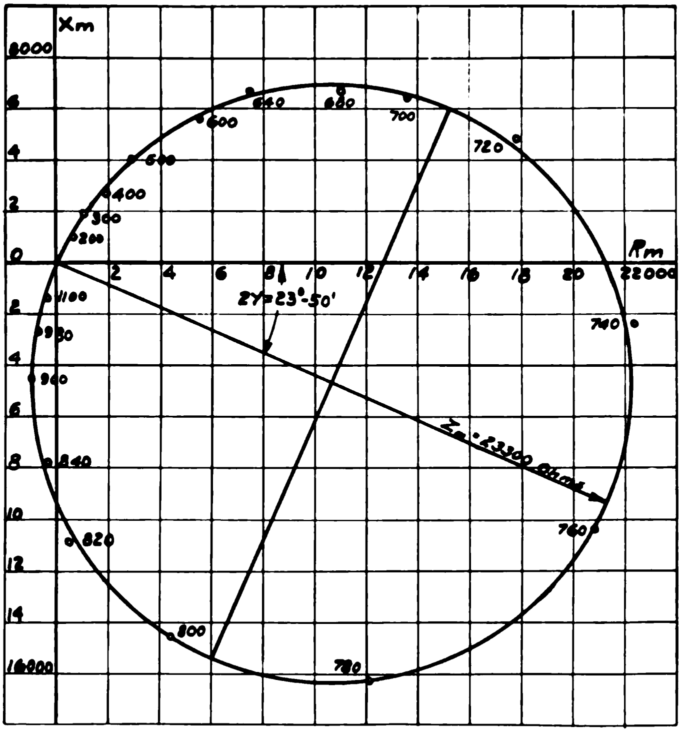

FIGURE 9—Motional Impedance Locus

of a diameter perpendicular to the principal diameter of the circle.

$$\frac{r}{2\times0.54}=\pi\cdot85$$
$$r=288$$

The force factor K of the receiver at the resonance frequeency is determined by the diameter Z_m of the motional impedance circle and the value of r.

$$Z_m=\frac{10^{-7}\,K^2}{r}$$
$$23{,}300=\frac{10^{-7}\,K^2}{288}$$
$$K=8.2\times10^6$$

K is a vector quantity which lags behind the current in the receiver windings by an angle γ equal to half the depression angle of the motional impedance circle. The value of γ for the receiver under test is 11° 55′ at 760 cycles. The vector K is plotted in Figure 10.

Since K is reduced principally by eddy currents as the frequency is raised, its value for direct current may be obtained by noting the effect of eddy currents on the receiver resistance and reactance. As the frequency is raised, the effective resistance (diaphragm clamped) increases because of eddy current losses, and the reactance is less than it would be if the eddy currents did not flow. The angle of lag θ of the eddy currents behind the voltage which produces them is given by the anti-tangent of the ratio of this decrease in reactance to the increase in resistance.* The phase of the eddy currents is the same as that of the force which they produce. So if a line is drawn from the tip of the vector K at an angle θ with the vertical (which is the phase of the voltage causing the eddy currents since it is in quadrature with the main current), the direct current force factor is determined by the point where this line cuts the horizontal line.

In the receiver tested the decrease in reactance of the receiver is 5,700 ohms, and the increase in resistance 1,340 ohms at 760 cycles. θ is therefore the anti-tangent of 4.25 or 67°. K_o, the direct current force factor, is found to be 15.2×10^6 dynes per ampere. The locus of K as the frequency is varied is a semicircle, as shown in Figure 10, points corresponding to other frequencies being determined by drawing lines from K_o at an angle (with the vertical) corresponding with the phase angle of the eddy currents for each frequency. Several points on the semicircle are shown.

The results of the test will be tabulated below:

$f_r = 760$ cycles.
$m = 0.54$ gms.
$r = 288$ dynes per (cm./sec.)
$S = m\ (2\pi\ f_r)^2 = 12.3 \times 10^6$ dynes/cm.
$K = 8.2 \times 10^6$ dynes/amp. at 760 cycles.
$\gamma = 11°\ 55'$
$K_o = 15.2 \times 10^6$ dynes/amp. (d.c. force factor).
$L_o = 3.18$ henrys at low frequencies.

The receiver was dissembled and thicker spacers placed in it so as to increase the air gap to a value sufficiently great to prevent the magnetic field from affecting the diaphragm stiffness. By means of a motional impedance test, the resonance frequency was found to be 910 cycles instead of 760 cycles. The intrinsic diaphragm stiffness

$$S_o = 0.54\ (2\pi \cdot 910)^2 = 17.6 \times 10^6 \text{ dynes/amp.}$$

* See paper "Theory of Magneto Mechanical Systems," by R. L. Wegel, Journal A. I. E. E.," October, 1921.

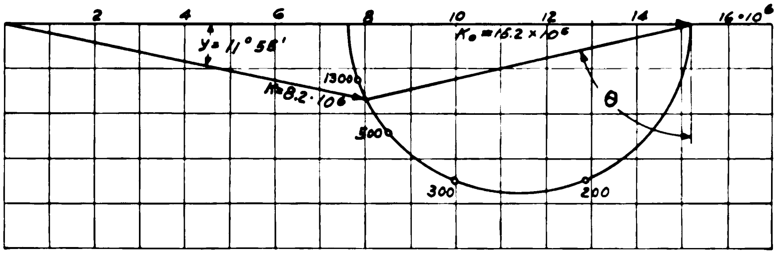

FIGURE 10—Force Factor Diagram

The reduction of rigidity due to the magnetic field in the case of the normal air gap is therefore

$$S_m = 17.6\times10^6 - 12.3\times10^6 = 5.3\times10^6$$

the predicted value

$$S_m = \frac{AM_o^2}{4\pi a^3} = \frac{0.189\times1072}{4\pi(0.0318)^3} = 5.4\times10^6$$

which is in close agreement with the experimental value.

The predicted value of the force factor for direct currents

$$K_o = \sqrt{10^{-7}L_o S_m} = \sqrt{10^{-7}\times3.18\times5.3\times10^{-6}}$$
$$= 13.0\times10^{-6} \text{ dynes/amp.}$$

This checks reasonably well with the value 15.2×10^6 determined experimentally

The force factor of this receiver is considerably greater than that of any simple receiver of equal inductance. This is accounted for by the balanced construction which allows the use of greater steady flux without diaphragm saturation.

VI. CONCLUSIONS

The balanced diaphragm type of receiver has all the advantages of the best types of electromagnetic receivers with the added feature that its vibrating system is just a simple diaphragm.

A determination of the characteristics of this type of receiver points to a direct method of design, with a degree of assurance that the greatest possible sensitivity will be obtained for a given application. The force factor, which is a measure of the sensitivity of the receiver, depends only on the inductance of the receiver, which is fixed by the characteristics of the tube circuit, and the allowable magnetic reduction of diaphragm rigidity, which is determined by the characteristics of the diaphragm.

The amount of power that a diaphragm can radiate without

serious distortion is roughly proportional to the cube of its area. Receivers with moderately large diaphragms are, therefore, capable of radiating sufficient power for great distances out of doors. Various sizes of receivers having equal inductance and fitted with horns which cause each to have the same uniformity of response to different frequencies, all have the same over-all sensitivity.

Research Laboratory, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania.

January 13, 1925.

SUMMARY: A discussion of the advantages and disadvantages of various present day electromagnetic receivers is given. A new type, called the balanced diaphragm receiver, is described and the details of design worked out.

AMPLIFICATION OF WEAK CURRENTS AND THEIR APPLICATION TO PHOTO-ELECTRIC CELLS*

By

G. Ferrié, R. Jouaust, and R. Mesny

(Paris, France)

In several instances, particularly in steller photometry, it is often necessary to measure very weak currents of the order of magnitude of 10^{-12} ampere. Galvanometers cannot detect currents of this nature, and consequently very sensitive electrometers are needed.

Tubes may be considered electrometers in which slight variations of the grid potential are manifested by considerable variations in the filament plate current. Thus the tube will act as a rugged and very sensitive electrometer which naturally can also be employed for measuring very weak currents.

Several attempts to develop this idea have been made by Kunz,[1] Pike,[2] and Meyer, Rosenberg and Lank.[3] Recently[4] we have applied the processes described below to the same problem.

A photo-electric cell constructed by Rougier[5] was employed.

This cell (Figure 1) is placed in a glass bulb having a diameter of 5 to 6 centimeters, equipped with tubulations for the passage of the wires which connect its electrodes to the voltage supply battery. The inside of the bulb is silvered, excepting for an aperture A which admits the luminous rays. On the silver plating opposite the aperture, at $K\,H$, is placed a hydride of potassium deposit electrically connected to the terminal C by means of a platinum wire. The anode P is ring-shaped in order not to inter-

*Received by the Editor, October 16, 1924. Translated from the French.

[1] J. Kunz: "Amplification of Photo-electric Current by an Audion," "Phys. Rev.," 10, 1917, page 205.

[2] C. E. Pike: "Amplification of Photo-electric Current by an Audion," "Phys. Rev.," 13, 1919, pages 102-108.

[3] E. Meyer, H. Rosenberg, and Lank (Zurich): "The Measurement of Photo-electric Currents by Means of Tube Amplifiers," "Arch. des Sc. Phys. et Nat.," 2, 1920, pages 260-262.

[4] G. Ferrié, R. Jouaust, R. Mesny: "Amplification of the Current of Photo-electric Cells and Its Use," "Comp. Rendue de l'Ac. des Sc.," 177, November 5, 1923, page 847.

[5] G. Rougier: "The Photo-electric Cells and Their Use for Photometry," "Revue d'opt. th. et exp.," 2, 1923, pages 133-166 and 365-383.

cept the light. The photograph (Figure 2) illustrates the dimensions of the cell.

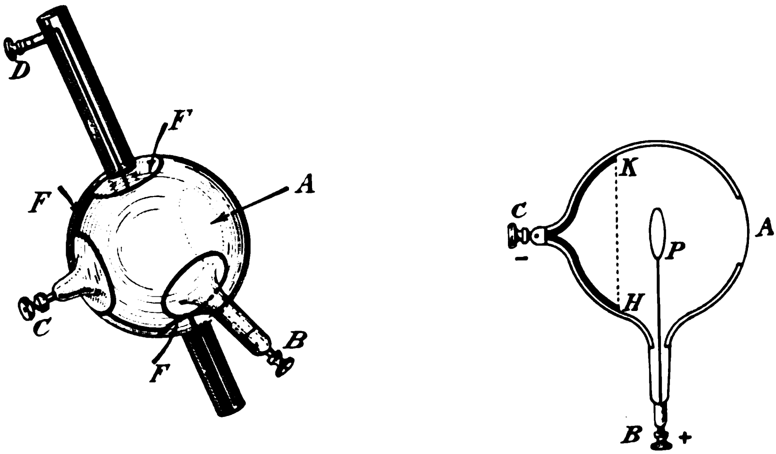

FIGURE 1

On the exterior surface are pasted sheets of tin foil *F*, serving as a guard ring for the prevention of weak currents due to the conductivity over the surface of the cell. These currents are superimposed on the photo-electric currents which are to be measured.

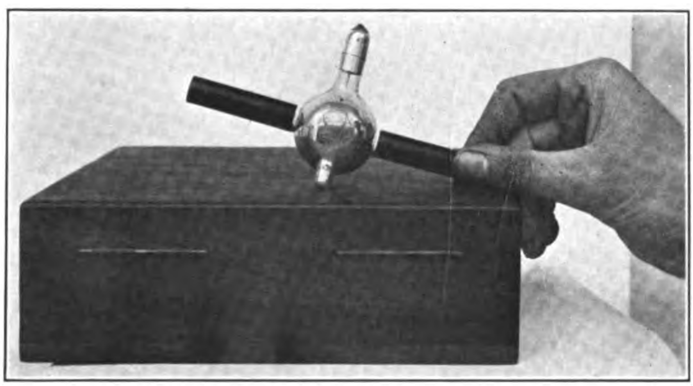

FIGURE 2

The connections which have been employed are as follows:

1. The anode of the cell is connected to the grid of a tube having three electrodes, the negative pole of the filament being connected to the positive pole of the battery, the negative pole

of which is connected to the alkaline deposit. As usual, a continuous difference in potential is applied between the plate and the filament of the tube (Figure 3).

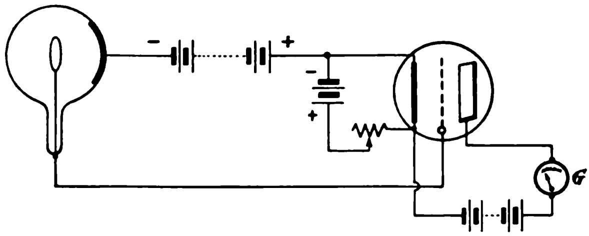

FIGURE 3

The emission of electrons, which takes place when the alkaline deposit is illuminated, gives a negative charge to the grid; as a result the filament plate current is reduced.

In order to obtain good results, the capacity between the grid and the various parts of the tube should be low. Moreover, since the grid must be highly insulated, it is well to employ tubes having "horns," and it is also necessary to select them carefully. Very few of the ordinary tubes are likely to give good results. With a tube having the dimensions of the ordinary receiving tubes we have obtained an amplification of 1,000, the variation of the filament-to-plate current being 1,000 times the intensity of the photo-electric current. An amplification of 10,000 was obtained with a 50-watt tube operating at 1,000 volts.

This connection is analogous to a method employed by Kunz and Rosenberg.

2. The difficulty in finding tubes with three electrodes having the necessary qualities has induced us to try another more complicated but also more reliable method (Figure 4).

A disc D, equipped with grooves, rotates at a great speed between the cell and the source of light. Thus the photo-electric current becomes a current of musical frequency corresponding to a rather high note.

The primary of the input transformer of an audio frequency amplifier, A, with three tubes is inserted in the circuit of the photo-electric current.

The amplified difference of potential, available at the output of this amplifier, is applied between the filament and the grid of a modulator tube M, in the plate circuit of which no steady electromotive force is inserted. But the plate-to-filament gap of this

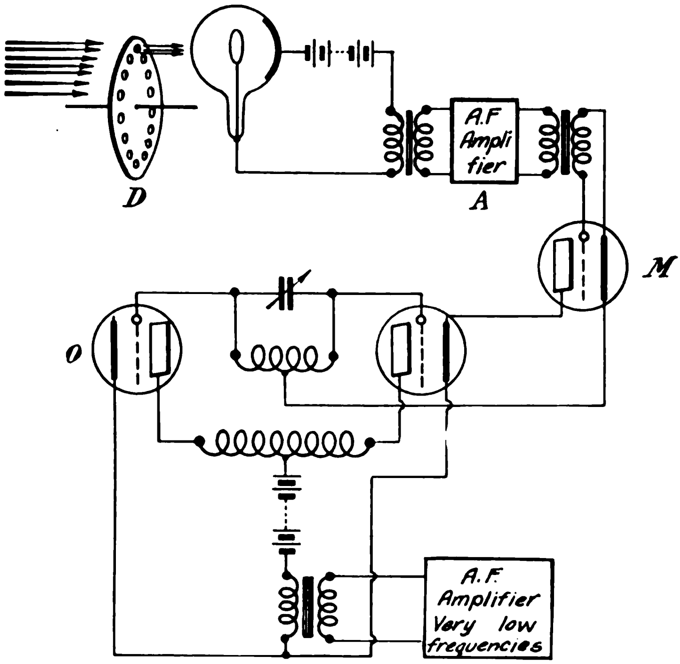

FIGURE 4

modulator tube is inserted in the grid-to-filament circuit of a small continuous wave generator *O*.

This generator was of the two-tube variety with a symmetrical connection (push-pull), and produced oscillations at a wave length of approximately 50 meters. When this apparatus is adjusted up to its maximum operating effectiveness, every variation of potential difference between the filament and the grid of the modulator tube (which causes variations of the filament-to-plate resistance of this tube) induces marked changes in the intensity of the oscillations produced; these variations in their turn produce variations in the plate-to-filament current of the oscillator. A further amplification can be obtained by conducting the latter variations to the terminals of an amplifier of very low frequency.

In this way amplifications of the order of 10^6 have been obtained.

By means of this process we have tried to register the photoelectric current due to the light of the stars; however, numerous difficulties are to be overcome in this respect. Nevertheless, experiments have been conducted under Jules Baillaud by using

the observatory of Paris equatorial telescope having an opening of 28 cm.

3. Finally, we have employed tubes with two grids in order to obtain a very powerful amplification with a single tube, thus eliminating the difficulty of selecting the tubes. For this reason we bore in mind the following considerations relative to tubes having only one grid:

When the cell is not illuminated, the grid collects a few of the electrons emitted by the filament. On the other hand, it receives a certain number of positive ions produced by the ionization due to the dissociation by impact of traces of gas remaining in the tube. Its potential, calculated with reference to that of the filament, has such a value that an equal supply of electricity is produced by the positive ions and by the electrons.

When the cell is illuminated, the electrons emanating from the cathode will charge the anode and the grid connected to it. The potential of the latter, which is already slightly lower than that of the filament, will be reduced; this reduction will necessarily increase the number of positive ions received by the grid; but the number of positive ions which can be received by the grid per unit of time is limited. If the illumination is intense, the number of electrons produced by the cathode of the cell per unit of time may be greater than this maximum number of positive ions. Then the negative charge of the grid will increase gradually, its potential will decrease and finally reach such a value that all electron emission from the filament is checked, that is, the plate current is interrupted.

If the illumination is weak, however, another state of equilibrium will take place such that the positive electricity supplied by the ions will be balanced by the negative electricity supplied by the electrons coming from the filament and the cell. In this way the plate current is reduced. Obviously, under these conditions any obstacle to the formation of positive ions in the tube will give greater effective significance to the variations of the plate current produced by a given illumination of the cell.

For this reason we have replaced the tube with three electrodes employed originally by a tube with two grids of the usual commercial type.

The exterior grid was connected to the anode of the cell, and a potential difference of 6 volts applied between the filament and the interior grid (Figure 5).

Under these conditions the interior grid assists in the emission of electrons. The voltage applied to the plate can be re-

duced, and was actually brought down to about 15 volts, a value slightly smaller than the ionization potentials of gases. In this way fewer positive ions were produced than in the ordinary tubes where the voltage between the filament and the plate was approximately 40 volts.

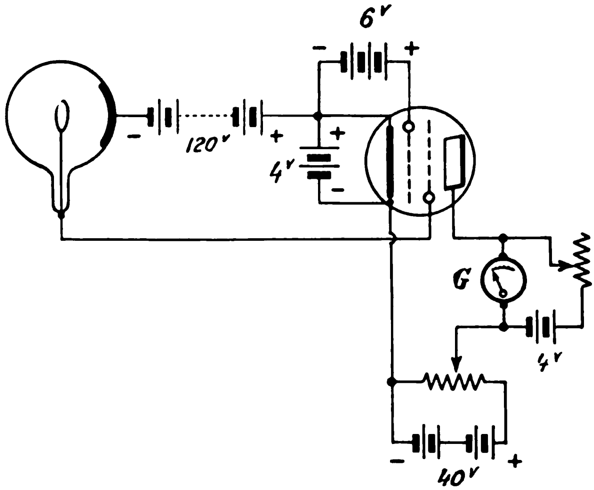

FIGURE 5

We have also proved that when the plate voltage of the tube is regulated accurately by means of a potentiometer, an amplification much superior to that obtained with the ordinarily better tubes can be produced. Any two-grid tube on the market can be used for this purpose.

We have looked into the possibility of applying this arrangement to stellar photometry, and with the aid of Jules Baillaud, experiments have been conducted at the observatory in Paris on the same equatorial telescope as mentioned above.

The cell, the tube with two grids, and the batteries which keep the first grid positive, were enclosed in a box covered by a grounded metal screen. This box was fastened to the photographic equatorial in such a way that the sensitive coating of the cell was slightly behind the focus. This box is shown in Figure 6 above the head of the observer. The arrangement has been chosen in order to make the luminous energy thus received operate on a large surface of potassium.

Carefully insulated wires were connected to the terminals of the filament heating storage batteries at convenient points, to the terminals of the batteries which were intended to act on

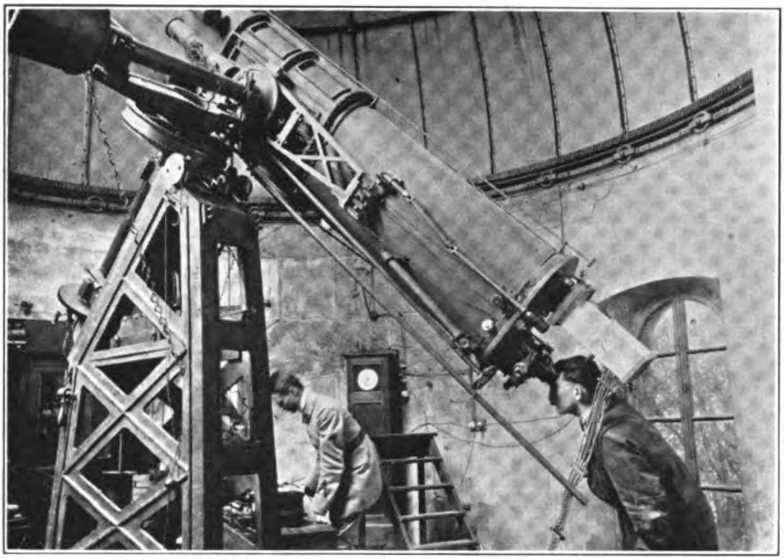

FIGURE 6

the cell, and to those batteries required to produce the necessary steady voltage in the tube between the filament on the one hand and the plate and the first grid on the other. A potentiometer made possible the regulation of this voltage. A sensitive galvanometer in the plate circuit of the tube was balanced for the normal plate current. The assembly of this apparatus is illustrated in Figure 7.

Under these conditions the star Capella gave a variation of the plate current of 3.5 micro-amperes, β of the constellation Bootis 1 micro-ampere, and ζ of the same constellation 0.3.

Let us consider the results obtained in previous experiments with rather casual arrangements. One might hope that an apparatus which is very carefully constructed with respect to insulation will have greater sensitiveness. This apparatus is now being built. It will make possible the direct utilization of the tube as an electrometer or as a ballistic galvanometer. It will also include a particular device to be explained in greater detail.

In view of what has been said above, it seems very difficult, despite the opinion of Rosenberg, to admit the proportionality between the variation of the plate current and the quantity of luminous energy received by the cell. Besides, even if this proportionality existed, the constant of proportionality would be subject to variations from one experiment to the other.

A standardization device has been provided, based on the

following principle: after each experiment a quantity of variable luminous energy is imparted to the cell. This quantity of luminous energy can be varied in a well-known way and regulated so as to obtain the same variation of the plate current as that produced by the observed star.

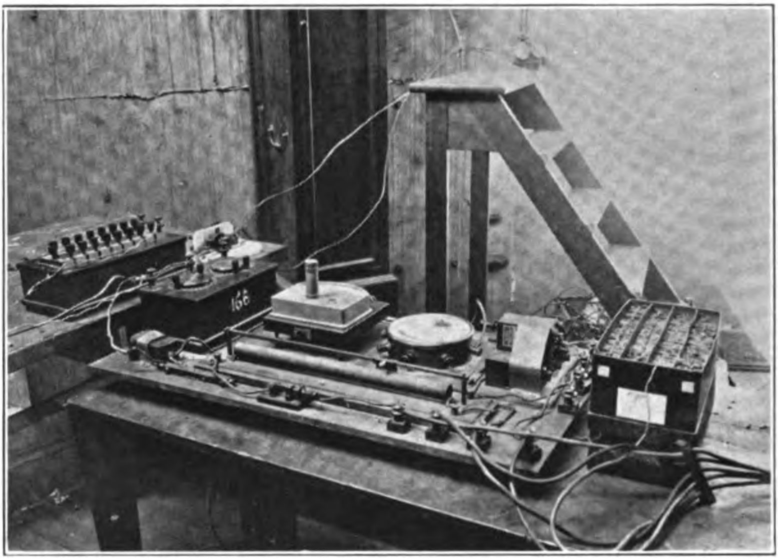

FIGURE 7

The practical utilization of this principle is easily understood: An electric lamp, the voltage of which is regulated accurately at its terminals will illuminate the part of the potassium layer which is influenced by the light of the star during an experiment.

Neutral glasses and absorbing glass wedges make it possible to vary continuously the luminous flux which arrives at the cell in this way.

4. In conclusion, another process should be mentioned which may yield a still greater sensitivity. A well-insulated condenser with a capacity of a few electrostatic units (c. g. s.) is placed in series with the cell, and the latter is subject to unknown illumination during a given time. Then the condenser is discharged by connecting one of its terminals to the filament and the other to the exterior grid of the tube (Figure 8).

An abrupt change in the plate current will then take place. In this way, by exposing the cell for 10 seconds to an illumination which can produce a permanent variation of the plate cur-

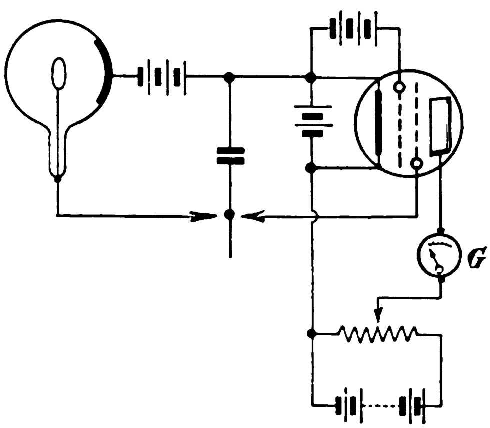

FIGURE 8

rent of 2 micro-amperes, we have obtained on the same measuring apparatus a deflection corresponding to a plate current change of 25 micro-amperes.

We will continue to study this new method of operation which also makes it possible to transform the light energy into telephone signals, when the luminous intensity is sufficient to allow a discharge of the condenser at least 15 or 20 times per second. It is only necessary to connect the arrangement in question to an ordinary amplifier for low frequency currents.

According to the above, it should be possible to use the properties of photo-electric currents for determining the instant when a star has a given position in the field of an instrument, and particularly the instant when it passes a meridian.

Naturally it may be possible to register on the same apparatus the time indicated by an astronomical pendulum and the photoelectric current produced by the observed star acting on the cell.

As long as the star is concealed by the micrometer, the photoelectric current will be reduced. Thus it is possible to observe the instant when the image of the star comes into contact with the micrometer cross thread, and the instant when it is entirely disclosed. Thus the personal equation, which must be considered in most meridian operations, is removed.

As a matter of fact, the practical realization of this new process presents very great difficulties which for the most part are not yet overcome. Nevertheless, we must mention certain previous rather encouraging experiments carried on at the observatory in Paris.

In order to get the desired results, it is necessary to measure the amplified photo-electric current with rapidly responsive indicating apparatus. Now such apparatus in general is only slightly sensitive. Therefore, the arrangement used for stellar photometry has been somewhat modified. As mentioned above, the current was amplified by means of a two-grid tube. A resistance of 50,000 ohms was inserted into the plate circuit of this tube, and the variations of voltage at the terminals of this resistance amplified by means of a new amplifying tube using direct current.

A Dufour galvanometer with photographic recording equipment was placed in the plate circuit of this new tube, the apparatus being such as is used by the geographic service of the army for range-finding by means of sound.

A plate having an aperture of one millimeter was placed in front of the cell in the photographic equatorial telescope previously mentioned. When the equatorial is held stationary, no photo-electric current is produced until the star, as a result of its apparent movement, passes in front of the aperture. In observing the star Vega, a deviation of a centimeter was revealed on the film of the photographic recorder, corresponding to the production of a photo-electric current having a certain duration of time, and the beginning and end of which seem to be shown accurately by observing a few precautions.

Altho only preliminary research has been conducted thus far, it seems possible that a new field has been found for photo-electric cells in astronomy.

Another application of the amplification of very weak currents has been studied by Lejay, who has employed it for measuring the potential gradient of the atmosphere by connecting the control grid to a potential terminal or test point. He has proven by comparison with a Mascart electrometer having photographic recording equipment that the errors amounted at most to two hundredths of the total deflections. He employed the ballastic method involving the changing of an auxiliary condenser.[6]

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[6] P. Lejay: "An Electrometer Using Triode Tubes and Its Use for Measuring the Electric Gradient of the Atmosphere," "Comptes rendus de l'Ac. des Sc.," volume 178, pages 1480-1482, April 28, 1924. "The Use of Tubes with Several Electrodes in Photometry," "C. R.," volume 178, pages 2171-2173, June 23, 1924.

GENERATION OF POLYPHASE OSCILLATIONS BY MEANS OF ELECTRON TUBES*

By

René Mesny

(Professor of Hydrography at the Laboratory for the Military Radio Telegraphy)

General Principle

If a connection having a geometrical and electrical symmetry of the order n is obtained by n identical triodes, the system constituted in this way should form under favorable conditions a unit of polyphase oscillations of the order n. This proposition is obvious; but it remains to be established *a priori*, whether the operation of such a system is stable. The slightest difference between two homologous elements, for instance, might interrupt the polyphase operation, thus producing several oscillations of different frequencies in the circuits. It seems to be impossible to verify the stability by means of calculation. Without considering the difficulties arising from the large number of elements in operation, too little is known of the phenomena taking place in the triodes. In contrast to this, a quick and safe method has been obtained experimentally.

The connection illustrated in Figure 1 was devised for three tubes, and the three-phase operation easily secured. The three filaments are in parallel from the same source; the three plates are connected at a common point P by means of three inductances, the same is the case with the three grids, the coils of which have a common point C; the coils of the plate and of the grid of the same triode are coupled magnetically. An electromotive force of some hundred volts is inserted between the point P and one of the terminals of the filaments, as for a single tube connection; the point C is connected to one of the filament terminals either directly or thru a resistance of approximately 10,000 ohms. The purpose of this resistance is to avoid overload of the grids by reducing very considerably the electron current which flows thru them; it is unnecessary to shunt it by a capacity as in the single triode connections because it does not transmit alternating current.

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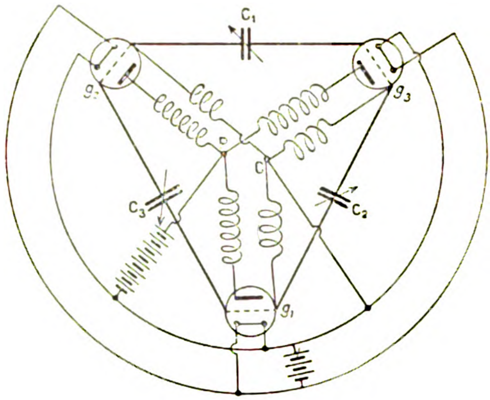


FIGURE 1

A variable condenser which produces an oscillatory circuit in conjunction with the corresponding inductances is placed between the grids (or the plates) of two adjacent tubes.

This system oscillates spontaneously in three phases, when the three oscillatory circuits, connected in this way have the same frequency, and the mutual inductances are suitably chosen.

The oscillations thus obtained are very stable and persist even with differences between the homologous elements amounting to 4 or 5 percent, a fact which makes it possible to carry different loads in the different circuits. When the differences between these elements are too large, the system develops oscillations at two or three principal frequencies, but the three-phase oscillations of a single frequency occur abruptly as soon as the limits indicated above are reached, as the capacities or the inductances are varied.

The unit can be considered to be the summation of three oscillatory circuits $g_1\ C\ g_2\ C_3$; $g_2\ C\ g_3\ C_1$, and $g_3\ C\ g_1\ C_2$, in part superimposed on each other, and each of them giving rise to an oscillation 120° behind its predecessor and 120° in advance of its successor or *vice versa*. The frequency is determined by the common values of the inductances and the capacity of each of these circuits to the same degree of approximation as for the single tube connections. Frequencies of the order of 10^6 cycles per second have been obtained without difficulty.

The existence of three-phase oscillations can be demonstrated by developing a rotary field by means of the same processes as those used for industrial frequencies, for instance, by sending the three currents 120° out of phase into three coils which are

themselves placed at an angle of 120°; these coils are inserted between the point C on the one hand and the three inductances connected to the three grids.

A squirrel cage motor, consisting of closed windings of fine wire and placed inside the coils mentioned above, begins to rotate rapidly as soon as the three-phase oscillations are generated. The rotation may be in either of the possible directions.

The power of the motor constructed in this way with frequencies of 5×10^5 cycles was of the order of 10^{-3} watts; however, by using a set oscillating on frequencies of 500 to 1,000 cycles per second, powers of a few watts could be obtained with a current of 1 ampere in the oscillatory circuits. It is probable that the speed of rotation in a vacuum would be very nearly constant. It depends only on the frequency of the oscillations, which can be kept almost constant, and on the mechanical resistances: if the motor rotates in a vacuum, the latter will be almost uniform, and the apparatus can be used for spinning mirrors at great speed and very regularly.

Measurement on the Separate Phases

The measurement which seems to be of greatest interest at high frequency, is that of the separate phases. In Figure 2, F_1 F_2 F_3 designate the three coils at 120° mentioned above. These coils have the axis O perpendicular to the plane of the figure and as a common axis of symmetry. The squirrel cage is replaced by a coil M, rotateable around the axis O, the turns of this coil being parallel to the axis of rotation. This coil will be the seat of an electromotive force of the same frequency as the oscillations and of a phase depending on the orientation of the movable coil. When properly constructed, the latter makes possible definite variations of the phase equal to the angle to which it is turned. We will assume that this condition, which will be considered later, is realized.

When the coil M is introduced in a circuit E C including a condenser C and another coil E, the entire circuit can be turned to the frequency in question. A receiver system, connected to the terminals of the condenser, picks up the oscillations induced in the circuit. If the orientation of the coil M is changed, the sound remains the same since the rotating field is circular. But, if the coil E is coupled to a third coil B, thru which flows a current of the same frequency, a coupling value and an orientation of M will exist for which no audible sound is produced; at this instant the electromotive forces induced in M by the rotating

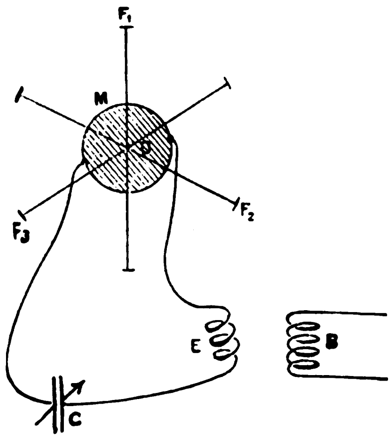


FIGURE 2

field and in E and B will be exactly opposite. A graduated circular scale is mounted on the axis O, its displacement in front of an index indicates the orientation of M and consequently the phase of the resulting current with respect to a phase used as an origin. In order to determine this origin, the point of extinction obtained by connecting E with a few turns in series with the circuit of one of the stationary coils, F_1, for instance, should be found. Zero is marked on the graduated circle at this point, and the figures read on the scale during the following measurements will constitute the differences in phase between the observed currents and the current in F_1.

The resulting points of extinction are very sharply defined and permit readings within a fraction of a degree.

The magnetic coupling between B and E can be replaced by a resistance and capacity coupling.

Naturally the frequency in the coil B should be equal to that of the generator of polyphase oscillations. This condition is easily obtained by producing the oscillations of the system which contains B by the generator itself. If, for instance, it is desirable to study the variations of the phase which occurs at the different stages of an amplifier using triodes, this amplifier is supplied by a potential difference in a section of the circuit F_1 which determines the reference phase.

In case the related phases in the different circuits of a generator are to be investigated, the three-phase system can be synchronized with the generator by establishing a suitable coupling between the latter and one of the three phases. When the

frequencies are adjusted to be sufficiently nearly alike, they drive one another and oscillate synchronously.

Accuracy of the Measurements

The first condition for obtaining the results below is the uniform proportional variation of the phase of the electromotive force induced in the coil M as a function of the angle thru which it is rotated. For this purpose it is sufficient to construct the coils F and M in such a way that the coefficient of mutual induction between M and the coil F_1 varies as the cosine of the angle between the planes of their windings. Let us call Φ, the flux produced in M by one of the coils F when their windings are parallel, and a the angle between the windings of F_1 and those of M for any position of the latter; then the entire flux which flows thru M in this position will be:

$$\Phi\left[\cos \omega t \cos a + \cos(\omega t + 120)\cos(a + 120) \right.$$
$$\left. + \cos(\omega t - 120)\cos(a - 120)\right]$$

the angle a being reckoned in the reverse direction to that of the variations of the phase between the coils F_1, F_2, F_3. This expression is reduced to

$$\frac{3}{2}\Phi \cos(\omega t - a)$$

which establishes the proposition.

On the other hand we have determined very general conditions for a mutual induction proportional to $\cos a$; they are as follows:

(a) The coil F should have a symmetrical plane passing thru the axis of rotation of M.

(b) The coil M should be wound on a core having the shape of a solid of revolution around its axis of rotation.

(c) The planes of the windings of M should be equally spaced.[1]

So far we have assumed that the currents circulating in F_1, F_2, F_3, have exactly the same intensity. Experience shows that polyphase oscillations can take place also when this condition is not fulfilled, if the threefold symmetry is imperfect, which always seems to be the case. Then the rotating field is elliptical, and variations in the sound heard are perceived accompanying a rotation thru 360° of the coil M. This defect can be corrected by adjusting the different elements of the circuits: couplings, capacities, resistances. If the two axis of the ellipse

[1] R. Mesny, "Radiation Measurements," "Onde Electrique," volume 1 (1922), pages 54-62.

of the rotating field are called a and b, the ear can easily perceive such deviation as

$$\frac{a-b}{a}$$

amounting to 1/10. Consequently, the flattening of the ellipse can be corrected without difficulty within the above limits. If more accuracy is desired, a voltmeter amplifier can be used instead of the ear.

On the other hand, when the field is elliptical it is easy to prove that the angle α, read on the scale, or the phase of the electromotive force induced in M for an orientation α of this coil, causes an error, at most equal to the angle ε given by the formula

$$\sin \varepsilon = 2\frac{a-b}{a+b}.$$

It should be added that the system of coils F and M of the generator should be spaced carefully in order to avoid parasitic inductions; a distance of 1.50 to 2 meters is suitable. It is also advantageous to inclose this system (F, M) in a cage of wire gauze.

Radiation of a Rotary Field

A rotary field can easily be radiated with this connection. It is sufficient to substitute for the coils vertical loops placed at 120° to one another.

Under these conditions, the radiated field[2] is symmetrical about the vertical axis passing thru the central point of the transmitter. The vector field describes an ellipse, the plane of which is perpendicular to the direction of the transmitter.

The small axis of this ellipse intersects the vertical axis of the field, its longer axis being horizontal; the ratio of the two is equal to $\cos \theta$, where θ designates the angle from the zenith of the point in question, as seen from the transmitter.

As one special case, the vector field describes a circle around the vertical axis of the system, and is polarized horizontally near the ground.

Such a transmission might serve an airplane for determining the direction in space of the transmitter and, consequently, for assisting in landing at a given point.

The word "field" is here conceived in the sense of the space where the radiation takes place.

THE SHIELDING OF ELECTRIC AND MAGNETIC FIELDS*

By

John H. Morecroft and Alva Turner

(Columbia University, New York)

An examination of the past Proceedings of this Institute shows no papers on the question of shielding; as the subject is becoming of increasing importance in radio apparatus it seems worth while to bring up a discussion of the matter at this time. The theory and experiments reported in this paper make no pretense at completeness, but they do serve, however, to point out some of the essential principles and to give data which apparently is not available elsewhere.

The general question of shielding naturally falls into two general classes: shielding against steady or constant fields and shielding against changing fields. The former is comparatively simple, from both theoretical and experimental viewpoints, whereas the latter can in general not be handled from the theoretical viewpoint except in an approximate manner. It is here, therefore, that experimental evidence is most needed and most of the experimental work following falls into this class.

Constant Magnetic Fields

Shielding against constant magnetic fields falls into two classes according to the way in which the field is set up; if by a permanent magnet one treatment is possible, whereas an electromagnetic field may demand entirely different treatment.

The flux from a permanent magnet is fixed in quantity; it can be neither increased nor decreased by changes in the surrounding conditions. Advantage is taken of this fact in the general scheme of shielding; the space to be shielded may be surrounded by a shield of highly permeable material (iron) and most of the flux will be diverted from the space to be shielded. Figure 1 shows the flux distribution after introducing an iron shell into the field of a permanent magnet. The shielding is not perfect

* Received by the Editor, December 16, 1924. Presented before The Institute of Radio Engineers, New York, December 17, 1924.

because any two points in the iron shell (A and B) must be at a difference of magnetic potential equal to the product of the flux and the reluctance of the intervening magnetic path. Because of this difference of potential there will be some flux from A to B. The amount of this flux can be kept low by reducing the magnetic potential between A and B, and this in turn is kept low by using a thick iron shield, or still better, a series of shields are inside the other.[1]

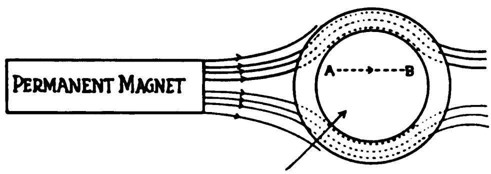


FIGURE 1

Where a comparatively large space is to be shielded from a uniform magnetic field such as that of the earth, Helmholtz's coils accomplish the purpose very well. The space to be shielded from the parallel field is indicated in Figure 2. Two large circular coils, A and B are placed co-axially, and their axis coincides with the direction of the earth's field; by sending a sufficient current, in the right direction, thru these coils the desired shielding effect is obtained. It is comparatively easy to reduce the earth's field by this method to less than 1 percent of its normal value.

If a conductor is carrying current a flux will be set up around the conductor. The amount of this flux is not fixed (for a given current), as was the case for the permanent magnet, but is determined by the reluctance of the magnetic circuit surrounding the conductor. If the space to be shielded is at A (Figure 3), a surrounding iron shell will accomplish the purpose; by making the shell sufficiently thick, any desired degree of shielding may be obtained. It might seem that the space A could be shielded by putting a heavy iron pipe around the conductor as indicated in Figure 4; one might think that all the flux which the conductor generates would go thru the low reluctance pipe leaving no flux to reach out in the space A. Unfortunately, such a method of shielding fails completely; there is just as much flux density at A with the iron pipe around the wire as when it is not there.

If the conductor is located centrally in the iron pipe, the flux

[1] "Phy. Rev.," volume IX, number 4, Wills, "On the Magnetic Shielding Effect of Tri-lammelar Spherical and Cylindrical Shells."

density in the air (both inside and outside the pipe) is exactly as it was without the pipe. The total flux surrounding the wire is increased because of the high permeability of the iron pipe, but the density of flux in the air remains unchanged.

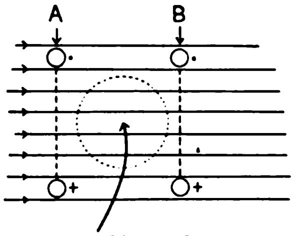


FIGURE 2

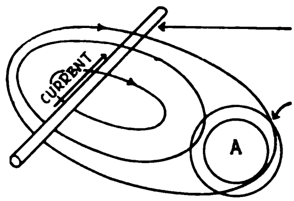


FIGURE 3

A general idea of the shielding accomplished by iron sheets in the form of pipe or otherwise may be gained from Figure 5. At any place +, close to the iron sheet, the flux density may be estimated by figuring the magnetic potential difference between two neighboring points on the shield, A and B. By supposing a Faraday tube to originate at A, go thru x, and end on B, the reluctance can be figured and the density at x then calculated. To shield the point x thoroly the two points A and B must be brought to a small potential difference, by some means or other. When an iron pipe completely surrounds the wire, the potential difference between points A and B is the same as if the pipe were not present. By shaping the iron in such a way that most of the magnetomotive force, due to the current, is used at the left side of the wire (Figure 5), shielding to some extent on the right side is made feasible.

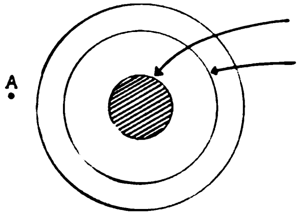


FIGURE 4

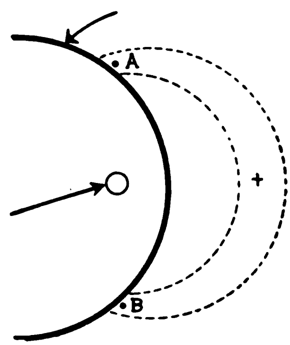


FIGURE 5

This effect is shown in Figure 6; a rectangular piece of sheet iron 0.11 centimeters thick was bent around the wire carrying the current, as shown in the lower part of the diagram. The flux at the point where shielding was desired was measured by a small search coil (2 centimeters in diameter), connected to a ballistic galvanometer. Twenty wires each carrying the same current were used to set up the magnetic field, the number of ampere conductors used being as high as 220. After the sheet of iron had been demagnetized the current in the wires was increased in steps, the field density was measured for each current value, and then the current was decreased in corresponding steps and the flux density again measured. This process was carried out both with the iron present and without it. The shielding is here defined as the ratio of the change in field strength, due to the presence of the iron, to the field strength with no iron present. If the iron so acts as to give no field at all at the point investigated, the shielding is perfect, or 100 percent.

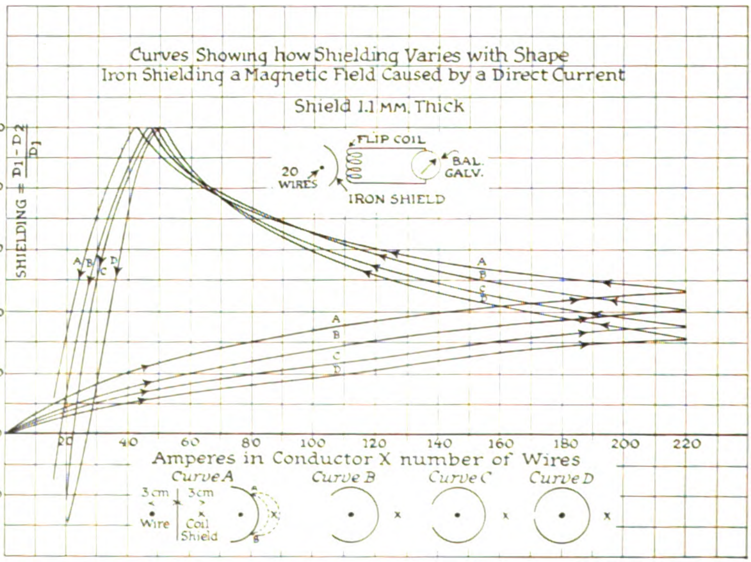


FIGURE 6

The shape of these curves (Figure 6) with increasing current indicates that the iron shield should be flat; a given length of iron shields better when plane than when partially surrounding a conductor. With decreasing current, the shielding increases because of the retentivity of the iron; for small values of current

(decreasing) the residual magnetism in the iron is sufficient completely to neutralize the field of the currents and perfect shielding results. For currents smaller than these, the ballistic galvanometer indicates that the flux present at the flip coil was in the opposite direction to that set up by the current, and for very weak currents the shielding becomes negative. There is actually more flux present with the iron shield than would be there if the iron were not present. The amount of this effect would quite evidently depend upon the magnetic properties (principally retentivity) of the iron used for shielding.

We would naturally expect that the shielding obtained in Figure 6 would be increased if the shield were made thicker; such is shown to be the fact in Figure 7, where three thicknesses of iron were used, the shield being in the form of a flat plate. Examination of the "decreasing current" part of these curves shows that the density of flux set up in the shield is greater for the thin shield than for the thick; the total flux thru the thick shield is greater but the density of flux is less.

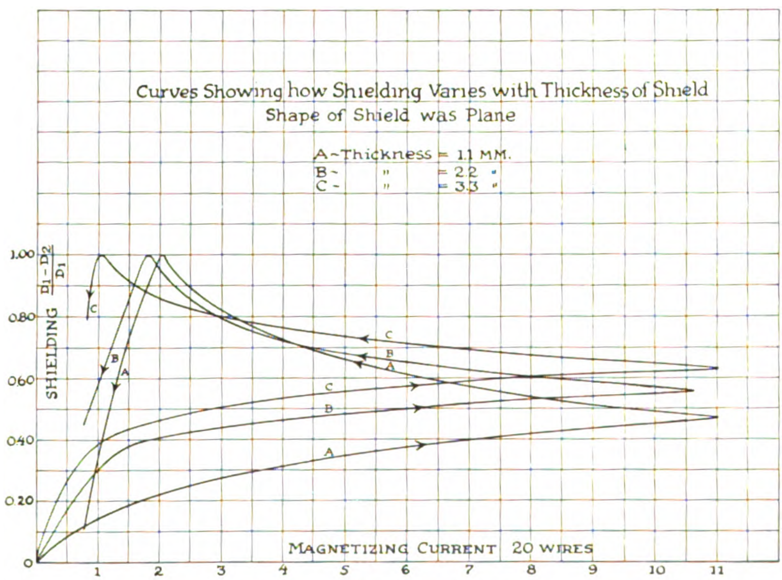


FIGURE 7

CONSTANT ELECTRIC FIELDS

The space surrounding an electric charge exhibits a radial field distribution, if the charge is isolated from other bodies; with increasing distance from the charge, the electric potential due to

the charge in question continually diminishes. Any other electric charge brought into the electric field will be urged to move towards or away from the first charge, according to its polarity.

Space in the proximity of an electric charge can be completely shielded from its field by surrounding the space in question with a completely closed metal cage, as suggested in Figure 8. Induced charges will be set up on the outer surface of the metal cage with such density and distribution that the net electric field inside the cage is zero. Actually the space inside must be considered as influenced by both the original charge and the induced charges; the induced charges will always so arrange themselves that there is no electric field at all inside the cage. It is to be noted that whereas the shielding is perfect after the induced charges have taken up their final disposition, the enclosed space is not shielded while this rearrangement of charges is taking place.

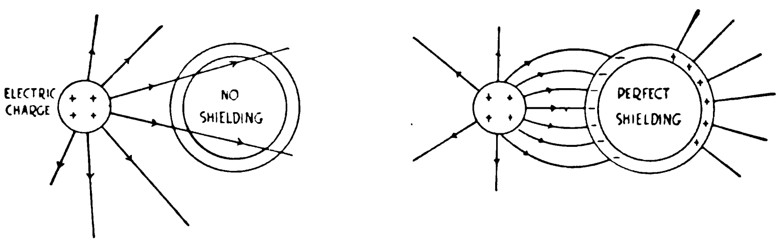


FIGURE 8

CHANGING MAGNETIC FIELDS

Consider a solenoid (Figure 9a) carrying a changing or alternating current. At any point A in the vicinity of the solenoid there is a magnetic field and, due to the rate of change of this magnetic field, an electric field. By interposing a sheet of metal between the solenoid and the point A (Figure 9b), the magnetic field is so redistributed that practically none penetrates to A. In case the metal shield is of iron, the shielding action previously referred to (Figure 5) will occur and in addition there occurs another action tending to shield point A. In the metal sheet, whether ferromagnetic or not, there will be set up eddy currents due to the alternating magnetic field, and the magnetomotive force at point A will be the resultant due to the solenoid and that due to the eddy currents. With a reasonably thick metal sheet of sufficiently low resistivity, the magnetomotive force of the eddy currents will practically neutralize that of the solenoid, so that

A is practically free of magnetic field. Being free of magnetic field we may say in general that it is also free of the electric field, due to the rate of change of this magnetic field.

The degree of shielding affected by eddy currents will be examined more in detail later in the paper, and experimental proof of the theory given.

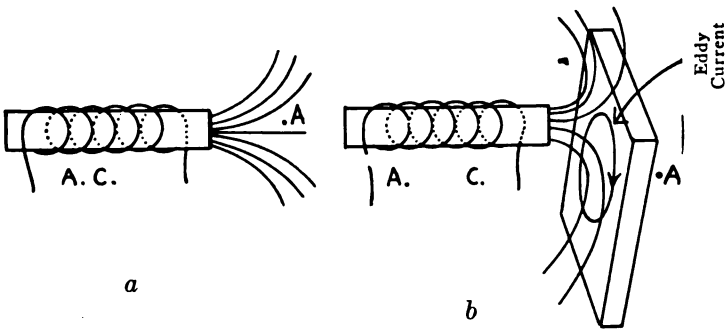


Figure 9

A very interesting illustration of apparent shielding is indicated in Figure 10; it represents a piece of sheath-covered wire, such as a submarine cable. If alternating current flows in the wire, voltages will be induced in the sheath and current will flow longitudinally, if possible. The formula for the mutual induction from wire to sheath is given in any handbook. The question may be asked—if alternating current flows in the sheath, will voltages be induced in the wire, these induced voltages, of course, being due to a changing magnetic field set up by the current in the sheath?

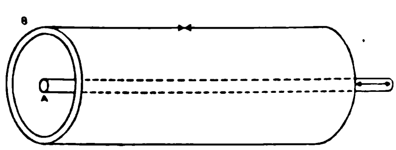


Figure 10

Now current flowing down a tubular conductor produces no magnetic field inside the tube, so that the apparently anomalous condition arises in which (if a voltage is induced in the wire) a voltage is induced by a changing magnetic field inside the tube when there is actually no magnetic field there. We must grant that voltage *will* be induced in the wire, because if there is mutual

induction between the wire and sheath, there must be a reciprocal action between the sheath and wire; mutual induction is not a one-way action.

A reasonably clear idea of the induction of voltage in the wire by alternating current in the sheath is obtained when we conceive of the electric current as the flow of electrons, which we now know it to be. It bothers the student very frequently in the study of induced voltages to get a concept of a changing magnetic field bringing about a motion of electrons (electric charges) when he has in mind the fundamental postulate that a magnetic field can exert no force whatsoever on an electric charge.

If we conceive of a magnetic field as nothing but an electric field in motion, the picture becomes much clearer; an electric field in motion *is* a magnetic field, the direction of which depends upon the direction of the electric field and the direction of its motion. With this picture in mind we can analyze the action which induces voltages in the wire of Figure 10.

Consider two oppositely placed filament elements of the tubular sheath, as indicated in Figure 11. Consider that the electron motion in the sheath is from left to right and that the electrons are accelerated in this direction. The fields of the electrons considered (*A* and *B* of Figure 11) are radial when they are stationary, or moving with constant velocity, but have backward "kinks" produced in them when the acceleration of the electrons takes place. These two kinks are shown in Figure 11 and it will be at once perceived that when such a kink traveling out from the electron with the velocity of light reaches the central wire, electron *C* in the wire will be urged from right to le t because of the component of electric field in this direction.

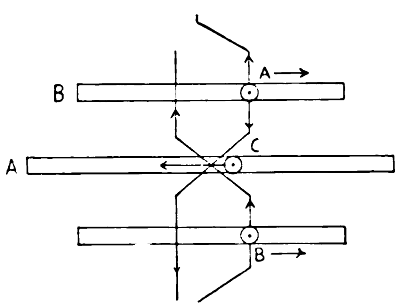


FIGURE 11

We know that induced currents always occur in opposite phase to that of the inducing current, and this picture at once gives us the reason therefor.

This same picture shows why there is no magnetic field inside a tubular conductor carrying current. All the elements of the tube may be considered in pairs, so that whatever action the pair of elements in Figure 11 brings about will be duplicated by all other pairs. The electric field from electron A, considered inside the tube, is down and moving to the right; it will produce a certain magnetic field. The electric field from B is upward and moving to the right; it will produce a magnetic field exactly equal to that produced by electron A, but in the opposite direction. The actual net magnetic field is, therefore, zero. Yet in ordinary nomenclature it is the rate of change of this field (of value constantly equal to zero) that induces the electromotive force in the wire of the cable.

Changing Electric Fields

In discussing the shielding of electric fields we said that any space completely surrounded by conducting material is completely shielded,—that the induced charges on the enclosing conductive cage completely neutralize the field which would exist without the cage. If, however, the electric charge, to which the field is due, is moving, the space inside the cage is not completely shielded. Thus in Figure 12 a charge is shown moving toward the shielding cage; insofar as the induced charges are not in their steady state (for the instantaneous position of the inducing charge under consideration), the space inside the cage is not shielded. The poorer the conductivity of the cage material, the poorer would be the shielding, because the induced charges would be so much the farther from their steady state disposition.

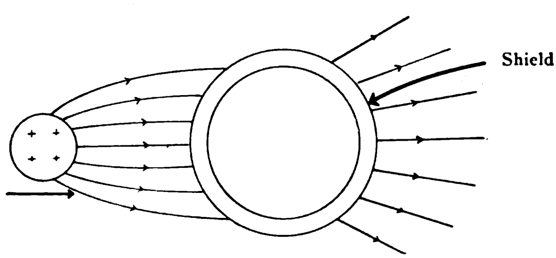


Figure 12

Leakage of Magnetic and Electric Fields

In general, a magnetic or electric field is desired in a certain part of space only; thus in a transformer it is desired that all the magnetic field be set up in the iron core where it is intended to go,

and in a condenser all the electric field is supposed to be directly between the two sets of plates. That these conditions are not so is shown by the induced voltage set up in a search coil in the vicinity of a transformer and by the change of note in an oscillating radio receiving set, for example, when the hand is brought into the vicinity of the tuning condenser.

An idea of the reason for these leakage fields may be obtained from the simple diagram shown in Figure 13; the coil sets up a magnetic field in the iron core and the magnetomotive force of the coil is used up thruout the different parts of the core (not only in that part of the core inside the coil). Knowing approximately the flux thru the core and the reluctance of the core, the difference in magnetic potential between two points A and B can be calculated. By imagining a Faraday tube between these two points, the flux in the tube can be calculated from the reluctance and the difference in magnetic potential between points A and B. This external or leakage flux will evidently increase with any factor which raises the difference in magnetic potential between A and B, such as increasing flux density or the presence of a secondary coil on the right leg of the core, this coil carrying a current opposite in phase to that of the current in the magnetizing coil. Such is the case of secondary and primary windings of an ordinary transformer.

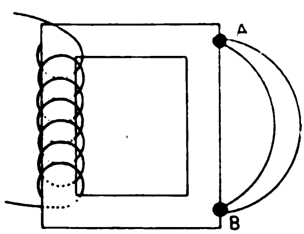


FIGURE 13

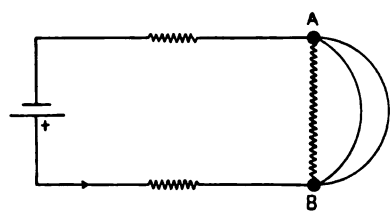


FIGURE 14

Electric leakage exists around an electric circuit for the same reason that magnetic leakage exists around a magnetic circuit, namely, a difference in electric potential between two points of the circuit. Thus in Figure 14 a current is made to flow around the circuit by the electromotive force of the battery; this emf. is used up all around the circuit, part of it in overcoming the resistance reaction between A and B. A hypothetical electric field circuit (a Faraday tube) joining A and B will have a field density proportional to the difference in electric potential of points A and B.

Circuits with No Leakage

It is frequently said that the toroidal form of coil has no magnetic leakage, that is, no external magnetic field; such a form of coil is shown in Figure 15, and we may readily see that there is no difference in magnetic potential between two points *A* and *B* picked at random on the periphery of the toroid. For if there is a difference in potential this difference must increase as we increase the distance between them, and by imagining the two points to separate farther and farther they will soon meet on the other side of the toroid. But when they meet there is evidently no difference in potential between them so we must conclude that that is never any difference in potential between them. If there is no difference in magnetic potential, however, between any two points *A* and *B*, there can be no magnetic field between them, and hence it is said that a toroid is a perfectly shielded magnetic circuit, having no external field.

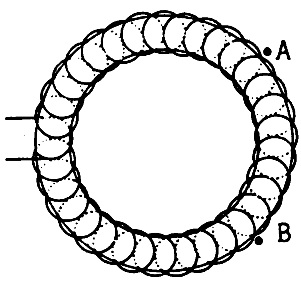


Figure 15

As a matter of fact this is not quite true; the toroid does have an external field and this is perpendicular to the plane of the toroid, that is, perpendicular to the paper in Figure 15. The amount of this magnetic fields is exactly the same as tho the toroid were a single turn of wire of the same diameter as the average diameter of the actual toroidal coil. This fact, while almost self-evident, was verified experimentally in our tests, by measuring the voltage induced in a search coil, first from a toroid and than from a single turn of the same diameter. The voltage induced in the search coil was the same for the two cases within experimental error.

An electric circuit around which there is no electric field (no electric leakage) is indicated in Figure 16a; a great number of cells are connected in series with an equal number of resistances to form a closed circuit. In such a circuit there is no difference

in electric potential between A and B and there will be no electric field between these points. Another circuit in which large currents may be flowing and yet have no measureable difference of potential between any two points is shown in Figure 16b.[2] An alternating magnetic field set up by a solenoid is supposed to thread a copper ring, the axis of the field being perpendicular to the paper (in Figure 16b) and thru the centre of the ring. Such an alternating field will set up large currents in the ring, yet any two points, such as A and B, have no difference of potential whatsoever between them.

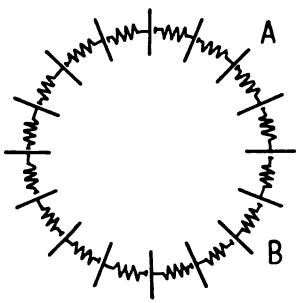


FIGURE 16a

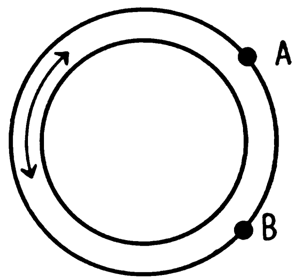


FIGURE 16b

The foregoing ideas regarding leakage fields may be summarized as follows—*any circuit in which the emf. (or mmf.) generated per differential length is the same as that used up in the same length, in overcoming the resulting reactions in this part of the circuit, will have no external field, as no point in the circuit will be at a potential different from that of any other.*

A Theory of Shielding

Leaving out of the question for the moment shields made of ferromagnetic materials, we may say that shielding is due to the eddy currents set up in the shielding material. In the general case, the paths of the eddy currents are not readily determined, so we will consider the easiest case, where the amount and position of the eddy current can be exactly determined. In Figure 17 is shown a short solenoidal coil A, another coil B, and between them a third coil C. If the current is set up in A, the magnetic field from this coil will naturally reach into B and induce voltages therein. This inducing field now has to link with coil C,

[2] In such circuits the potential is said to be a multiple valued function, having varying values as the circuit is traversed one or more times in the same direction.

however, and will set up currents therein which will tend to prevent the flux from A reaching into B. Coil C is said to be shielding coil B. The greater the eddy currents in C (up to a certain limit) and the more nearly they lag 180 degrees behind the current in A, the more perfectly is B shielded from A.

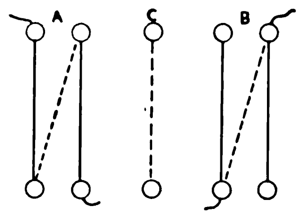


FIGURE 17

The extent of the shielding action can be derived as follows:

Let M = coefficient of mutual induction between A and B when C is absent (that is, without C). Then the voltage induced in B by a current i in A is given by $e = M\frac{di}{dt}$. Let M' = the correspondingly defined value of M when coil C is present. Then we shall define the shielding as equal to $\frac{M-M'}{M} \times 100$ percent. Evidently if the shielding is perfect M' = zero and the above formula yields a value of 100 percent.

To determine the shielding due to a certain arrangement, therefore, it is only necessary to measure the mutual induction with and without the shielding circuit present and to use the above formula.

The best method for measuring M will depend somewhat on the constants of the coils A and B and of the frequency to be used. At audible frequencies the best method is to measure the effective self-induction of the two coils A and B connected in series, once with the mmfs. in the same direction and again with the mmfs. in opposition. One-quarter of the difference of the two values for L so found is the mutual induction of the two coils, as can readily be seen by writing the equation of reactions of the circuit for the two cases.

Above audible frequency the bridge method may still be used employing the heterodyne method of determining when the bridge is balanced. At frequencies so high that unknown errors creep into the bridge determination, the resonance method may best be used for determining the self-induction required.

In case the shielding circuit involves iron, the bridge scheme is not applicable, as will be pointed out later.

A Short-Circuited Coil Used as Shield

Let two coils of inductances L_1 and L_2 be connected in series across a voltage $E_m \sin \omega t$, so that the field of L_2 can be reversed with respect to that of L_1 as shown in the diagram of Figure 18. Let a third coil L_3 be inserted between coils 1 and 2

The calculation of the change in mutual induction M_{1-2} can be most easily carried out by considering the induced voltage in the circuits. If it is supposed that one ampere (effective) of current is flowing in coils 1 and 2, then the induced voltage in this circuit, 90 degrees behind the current, is at once a measure of the reactance of the circuit. In part, this induced voltage is due to the self-induction of coils 1 and 2, and their mutual induction, while the rest of the induced voltage must be due to current set up in coil 3 reacting back on the original circuit.

If we consider two coils only, the second one being short-circuited, with a current of one ampere in coil 1, the induced voltage, E_2, in coil 2 will be ωM, and the current will be $\frac{\omega M}{Z_2}$ The component of this current 90° behind the voltage E_2 will be $\frac{\omega M}{Z_2^2} \omega L_2$. This current will induce in circuit 1 a voltage 90° behind itself in phase and in magnitude equal to $\frac{(\omega M)^2}{Z_2^2} \omega L_2$, and this voltage in circuit 1 will be in phase opposition to the inductance reaction in this circuit.

The total induced voltage in circuit 1, 90° behind the current in this circuit is $\omega L_1 - \frac{(\omega M)^2}{Z_2^2} \omega L_2$, and this is the effective reactance of circuit 1. The effective self-induction can be obtained by dividing thru this expression by ω. It will be seen that the presence of circuit 2 diminishes the self-induction of circuit 1 by an amount $\left(\frac{\omega M}{Z_2}\right)^2 L_2$.

The coils of Figure 18 can be treated in the same manner, remembering that the current in coil 3 is due to emfs. induced from both coils 1 and 2.

Let R_1, R_2, and R_3 be the resistances of the three coils and M_{1-2}, M_{1-2}', M_{2-3} and M_{1-3} the mutual inductances between coils 1 and 2, 2 and 3, and 1 and 3, M_{1-2} representing the case in which coil 3 is open-circuited and M_{1-2}' the case in which the circuit of the

coil is closed. Assume the reversing switch to be thrown so that the fields of coil 1 and 2 are in the same direction and the effective voltage E to be such that the effective current in coils 1 and 2 is one ampere when coil 3 is short-circuited. The effective induced voltage in coil 3 due to the changing current in coils 1 and 2, and 90 degrees behind the current will be

$$\omega\,(M_{1-3}+M_{2-3})$$

The component of the current in coil 3 due to this voltage and 90° behind it will be

$$\frac{\omega^2(M_{1-3}+M_{2-3})L_3}{Z_3^2}$$

The voltage in the circuit made up of coils 1 and 2, due to this component of current and 90° behind it, will be

$$\frac{\omega[\omega^2(M_{1-3}+M_{2-3})^2\,L_3]}{Z_3^2}$$

This voltage will be 270° behind the one ampere current supposed in the circuit of coils 1 and 2, and therefore 180° behind the inductive reaction $\omega(L_1+L_2+2\,M_{1-2})$. It can, therefore, be considered as an inductive reaction, and the effective inductive reaction in the circuit containing coils 1 and 2 is decreased by it to the value given by

$$\omega(L_1+L_2+2\,M_{1-2})-\frac{\omega[\omega^2(M_{1-3}+M_{2-3})^2L_3]}{Z_3^2}$$

The effective inductance in this circuit when coil 3 is short circuited is therefore

$$L'''=L_1+L_2+2\,M_{1-2}-\frac{\omega^2(M_{1-3}+M_{2-3})^2\,L_3}{Z_3^2} \qquad (1)$$

When the reversing switch is thrown so that the electromagnetic lines of force linking coil 2 are in the opposite direction to those linking coil 1, and coil 3 is short circuited, the voltage in coil 3 due to the one ampere current in coil 2 is opposite in phase to that due to this current in coil 1. The effective voltage in coil 3 due to one ampere in coil 1 and 2, and 90° behind it will be $\omega\,(M_{1-3}-M_{2-3})$. The component of the current in coil 3, 90 behind this voltage, will be

$$\frac{\omega^2(M_{1-3}-M_{2-3})L_3}{Z_3^2}$$

The voltage induced in coils 1 and 2 due to this component of current and 90° behind it will be

$$\frac{\omega[\omega^2(M_{1-3}-M_{2-3})^2\,L_3]}{Z_4^2}$$

As in the preceding case, this voltage will be 180° behind the inductive reaction and the effective inductance in this circuit when coil 3 is short-circuited is therefore

$$L^{IV}=L_1+L_2-2\,M_{1-2}-\frac{\omega^2(M_{1-3}-M_{2-3})^2\,L_3}{Z_3^2} \tag{2}$$

The effective mutual inductance between coils 1 and 2, M_{1-2}', is $\frac{L'''-L^{IV}}{4}$ and from equations (1) and (2)

$$M_{1-2}'=M_{1-2}-\frac{\omega\,L_3}{Z_3^2}\,M_{1-3}\,M_{2-3}$$

The change in the mutual inductance between coils 1 and 2, due to coil 3, which is short-circuited, is therefore

$$\Delta\,M_{1-2}=M_{1-2}-M_{1-2}'=\frac{\omega^2\,L_3}{Z_3^2}\,M_{1-3}\,M_{2-3} \tag{3}$$

Expressing the mutual inductances in terms of the coefficients of coupling of their respective coils, we have $M_{1-2}=k_{1-2}\sqrt{L_1\,L_2}$, $M_{1-3}=k_{1-3}\sqrt{L_1\,L_3}$, and $M_{2-3}=k_{2-3}\sqrt{L_2\,L_3}$. The expression we shall use as a measure of shielding is $\frac{\Delta\,M_{1-2}}{M_{1-2}}$ and from equation (3) and the preceding relations we have

$$\frac{\Delta\,M_{1-2}}{M_{1-2}}=K\frac{\omega^2\,L_3^2}{Z_3^2} \tag{4}$$

where $K=\frac{k_{1-3}\times k_{2-3}}{k_{1-2}}$ which is constant for a given arrangement of coils. Also we have $Z_3^2=R_3^2+(\omega\,L_3)^2$, and $\omega=2\,\pi f$.

The shielding depends on the frequency of the alternating electromagnetic field and on the inductance and resistance of the shielding coil according to the expression of equation (4). The effect of the frequency and resistance only were examined experimentally, as it is not feasible to vary the inductance, keeping the resistance constant. Theoretical values of $\frac{\Delta\,M_{1-2}}{M_{1-2}}$ were calculated from equation (4) using the same values of K, L_3, R_3, and f as were used to obtain the experimental results.

The coils used for the experimental results were wound with Number 24 insulated copper wire on hard rubber forms 1.3 cm. wide and 12 cm. in diameter. The coils used for frequencies up to 50,000 cycles were bank wound with 4 layers of wire having the respective inductances $L_1=0.777$ mh., $L_2=0.732$ mh., and $L_3=0.760$ mh. The shielding coil L_3 was used for all frequencies.

The two coils used with L_3 for frequencies between 50,000 and 100,000 cycles were bank wound with two layers of wire having the inductances $L_1 = 0.2160$ mh. and $L_2 = 0.2413$ mh. and the two used with L_3 for the high frequencies were single layer coils of inductances $L_1 = 0.075$ mh. and $L_2 = 0.077$ mh.

The three coils were arranged with the shielding coil, L_3, between the other two coils so that the coefficient of coupling between each pair was a maximum. One of the two outside coils was connected in series with the other by means of a reversing switch as shown in Figure 18. The shielding coil was connected in series with a variable resistance to determine the effect of resistance of the shielding circuit on the amount of shielding obtained. The curves of Figures 19 and 20 show the theoretical and the experimentally determined results; they differ by amounts within the experimental error.

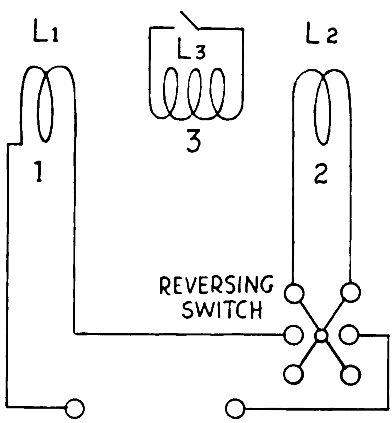


FIGURE 18

USE OF SHEETS OF NON-PERMEABLE METALS AS SHIELDS

The shielding coil, L_3, was replaced by square sheets of different thicknesses of copper, brass, and tin foil in this case.

In general, shielding is accomplished about as indicated in Figure 21; a sheet of conductor is interposed between the coil carrying current, A, and the coil B to be shielded from the current. The current in the shield in the case is in the form of a circular band having different densities in different parts of the shield about as indicated in Figure 21. The density will be greatest at the surface of the metal sheet next to coil A, and in this surface it will be greatest where it is closest to the conductors

of coil A. In Figure 21 an attempt to show this is made by the proximity of the + and − signs.

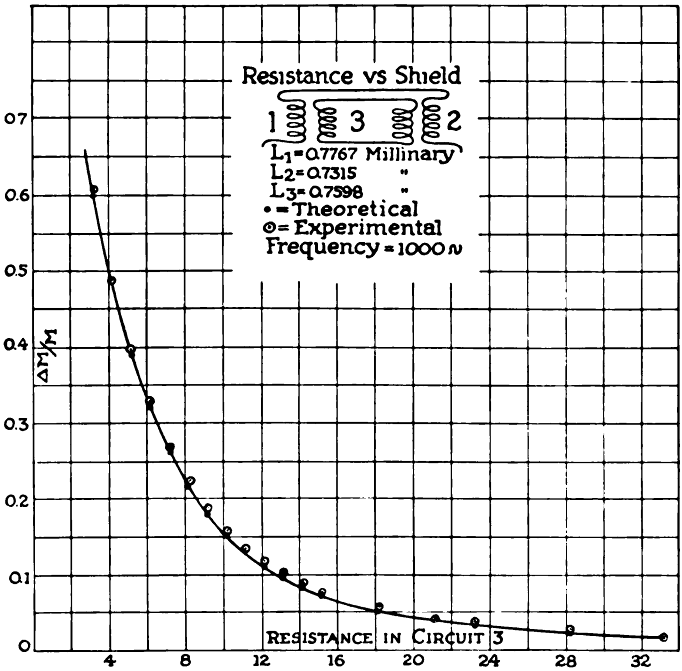


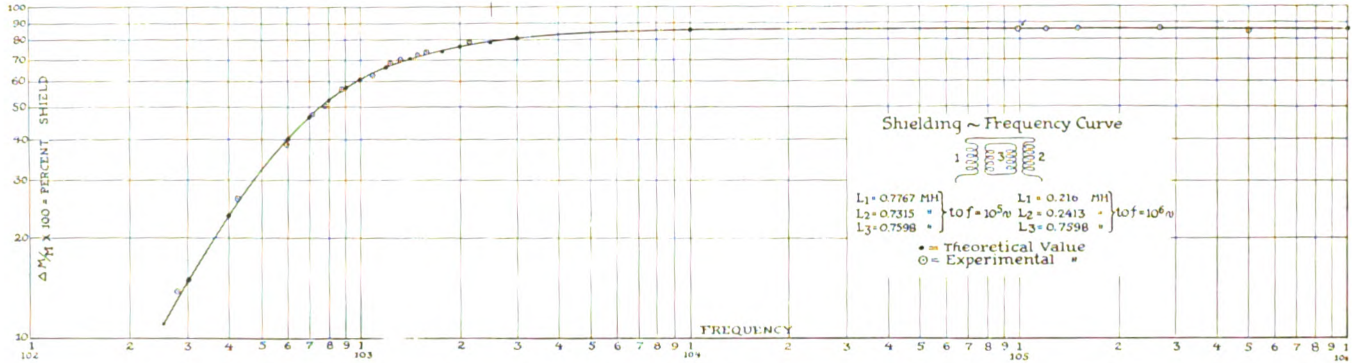
FIGURE 19

The curves of Figures 22, 23, and 24 give the relation between shielding and frequency for different thicknesses of the different metals between 0 and 18,000 cycles. The curves of Figures 25 and 26 show this relation for frequencies between 1,000 and 1,000,000 cycles. Logarithmic paper was used for these two sets of curves due to the large range of frequencies.

The shielding depends on the thickness of a particular kind of shield for a constant frequency as shown by the curves of Figures 27, 28, and 29.

The effective resistance of coils 1 and 2 depends somewhat on the frequency of the alternating electromagnetic field when the shield is between the coils and the fields of the coils are in the same direction. The curves of Figures 30, 31, and 32 show this relation for copper, brass, and tin foil as shields.

It is to be noticed that for a given frequency a definite thickness of shield introduced a maximum resistance into the shielded circuit.



Shielding ~ Frequency Curve
1
3
2
L1 = 0.7767 MH
L2 = 0.7315 "
L3 = 0.7598 "
to f = 10^5 ~
L1 = 0.216 MH
L2 = 0.2413 "
L3 = 0.7598 "
to f = 10^6 ~
• = Theoretical Value
⊙ = Experimental "
ΔM/M x 100 = PERCENT SHIELD
FREQUENCY

Figure 20

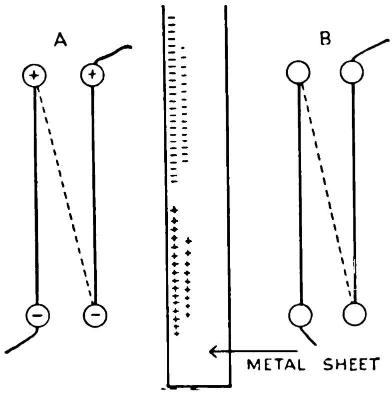


FIGURE 21

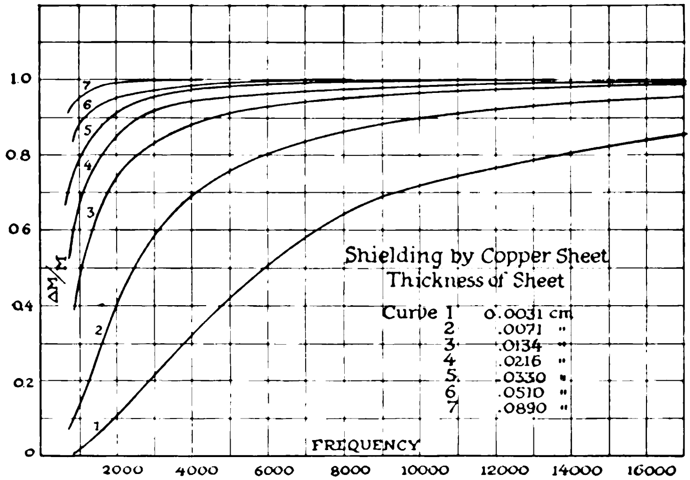


FIGURE 22

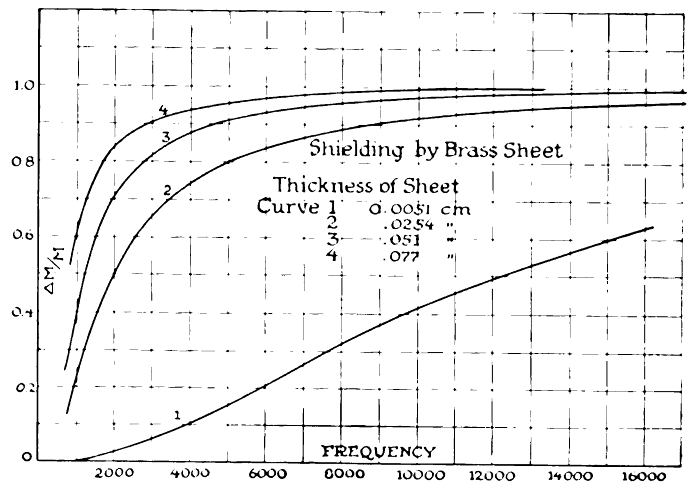


FIGURE 23

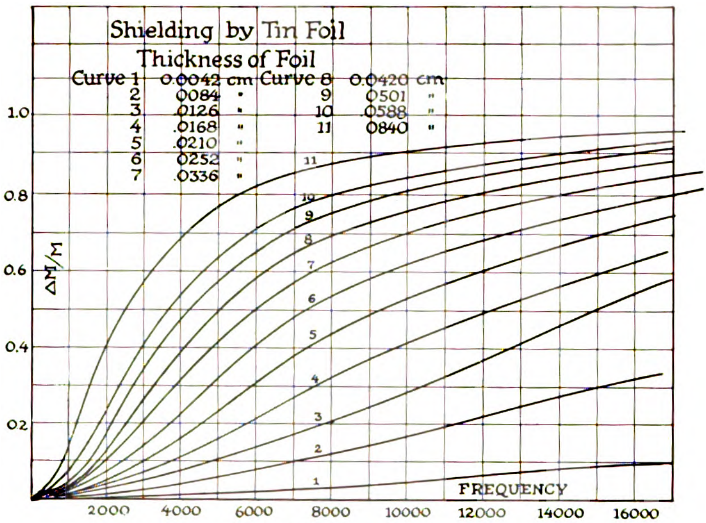


Figure 24

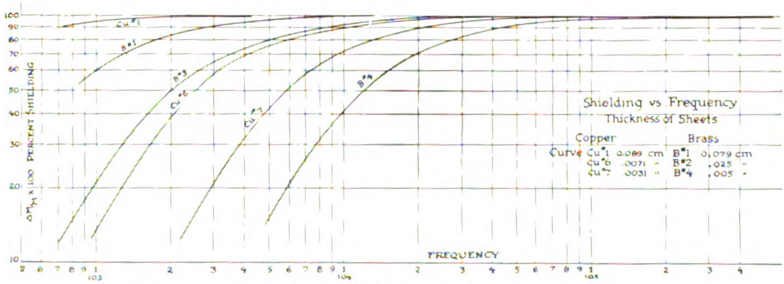


Figure 25

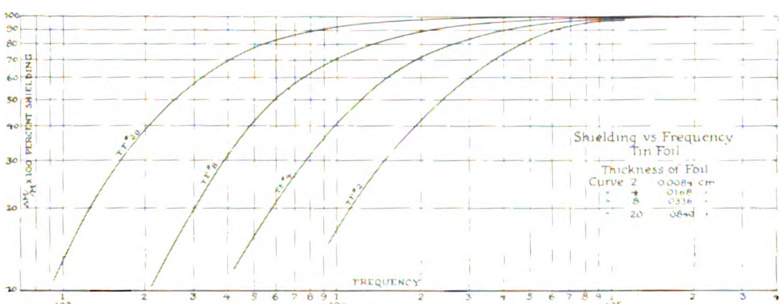


Figure 26

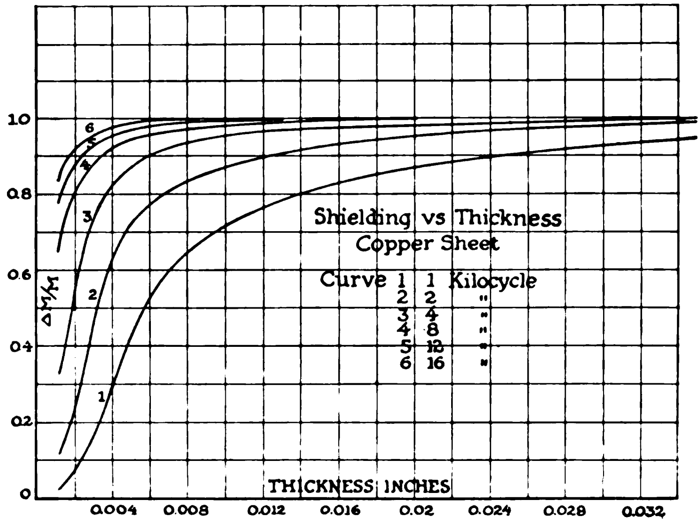


FIGURE 27

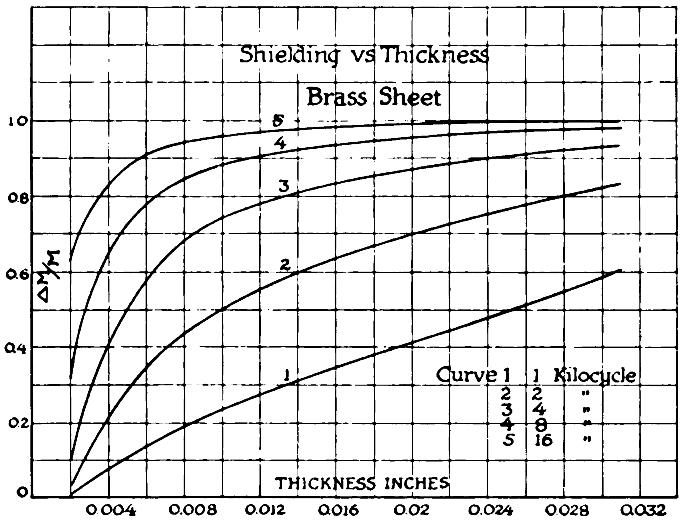


FIGURE 28

The shielding depends on the resistivity of the shielding metal for a definite frequency and thickness as shown by the curves of Figure 33.

The conclusion to be drawn from these results is that maximum shielding with minimum increased resistance is obtained by using a thick sheet of material of high conductivity.

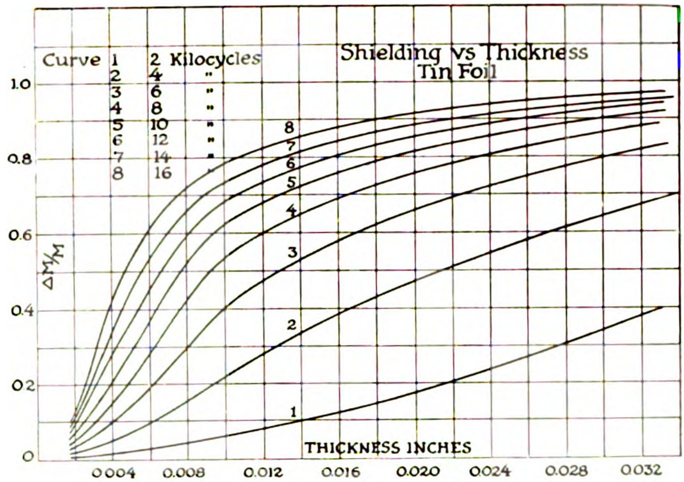


FIGURE 29

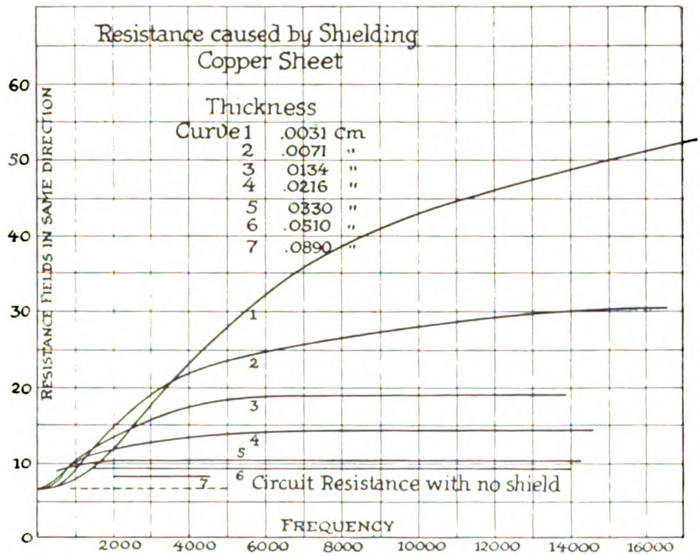


FIGURE 30

Position of Shielding Plate

The square sheet of metal used for a shield was arranged in each case so that its center coincided approximately with the intersection of the line joining the centers of the two coils and the plane of the shield. The shield was moved about, keeping the coils in the same position with respect to each other and no

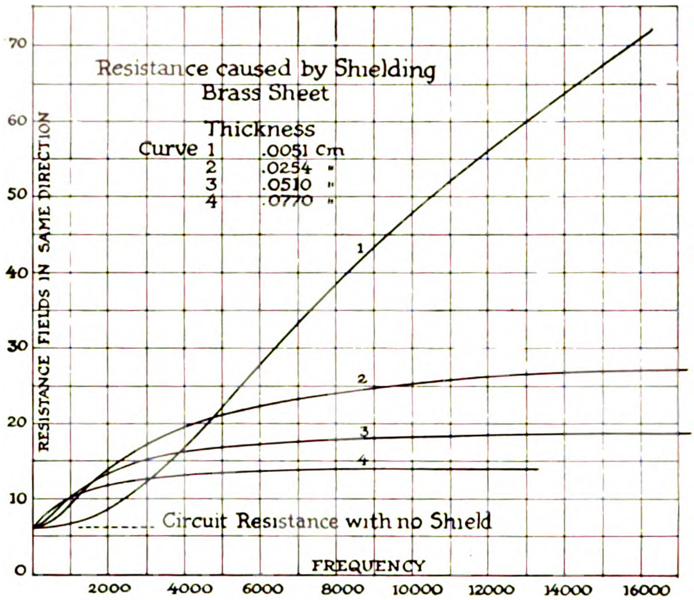


FIGURE 31

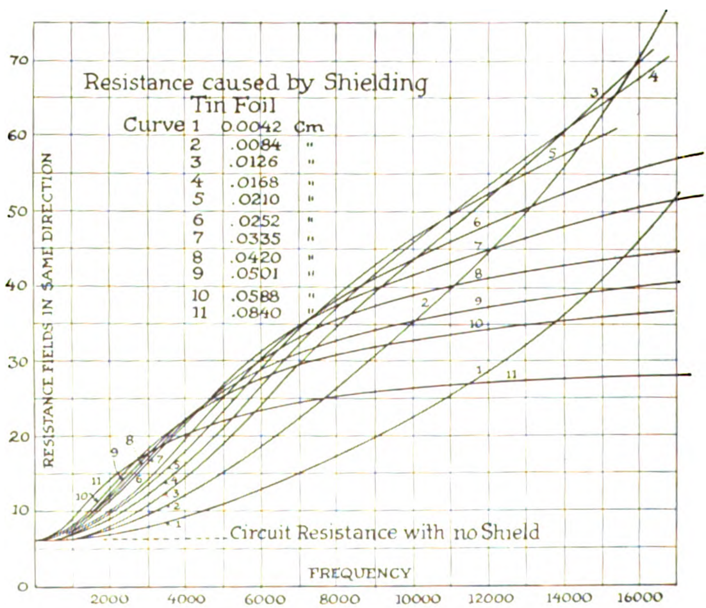


FIGURE 32

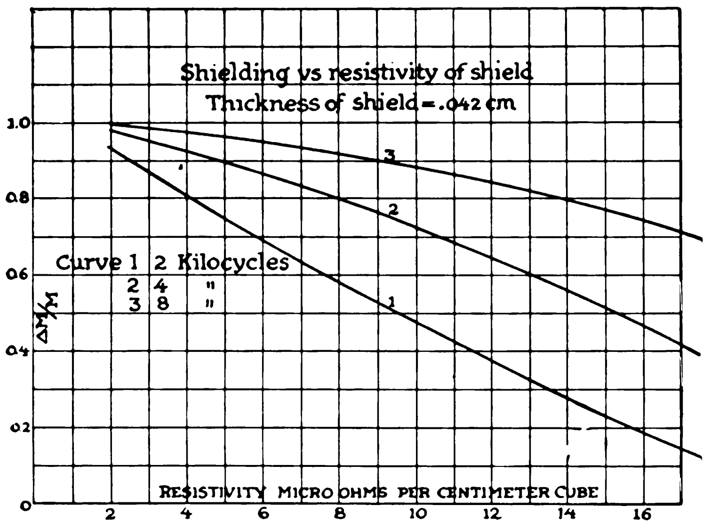


FIGURE 33

difference in the shielding effect was noticed until less than a half an inch of the shield was extending outside of the coils on one side. This was tried with copper and zinc and the results were the same.

EFFECT OF SHAPE OF SHIELDING PLATE

A sheet of copper 0.021 cm. thick was slit diametrically across the effective shielding area and $\frac{\Delta M_{1-2}}{M_{1-2}}$ was obtained for 1,000 cycles. The result was 4 percent less than that for the same sheet without the slit. A second diametrical slit was made in this sheet of copper at right angles to the first slit and $\frac{\Delta M_{1-2}}{\Delta M_{1-2}}$ was obtained for the same frequency, giving a result 4 percent less than that for one slit.

An ordinary piece of copper mesh was used as a shield for different frequencies, and the results showed very little shielding effect. A border of solder 0.5 cm. wide was put around the edge of this shield so as to make good contact between the ends of the wires, and the shielding was increased approximately 75 percent.

In one case where a high degree of shielding was necessary, all the apparatus was mounted on a heavy copper plate, and a heavy copper box arranged to lower over the apparatus, to rest on the copper plate. Altho the fit was reasonably good, the

shielding was comparatively poor. Some copper strips were soldered on the base plate of copper to make a ditch into which mercury could be poured; the ditch of mercury was so placed that the lower edge of the copper cover rested in the mercury. The shielding was now nearly perfect and the idea to be derived from this experience may be stated as follows: *if shielding is to be obtained by eddy currents, they must be free to flow as they will.* Any imperfect joint in the shield, which tends to constrain the eddy currents to restricted paths, will seriously interfere with the shielding obtainable.

The Case which Makes Use of Sheets of Iron as a Shield

When iron is used as the shield, the previously-described method of measuring the shielding leads to anomalous results; thus the two coils, connected in series, having an iron plate between them for shield may have an effective self-induction greater when their mmfs. are in opposition than when they are in conjunction.

To measure the shielding in such a case, the apparatus was arranged as in Figure 21; coil A is connected to a source of variable frequency thru an ammeter and coil B is connected to a vacuum tube voltmeter. The mutual induction between the two is proportional to the voltage induced in coil B, the current and frequency used in coil A. Now the current and frequency of coil A being maintained constant, the iron shield was slipped between the two coils and the voltage induced in coil B was again measured. If the shielding were perfect, no voltage would be induced in B; the shielding is given by the ratio of the difference of the two voltages to the original voltage.

As was the case for other shielding metals, it would be expected that the shielding would increase in amount as the iron was increased in thickness; such is shown to be the case in Figure 34. At the lower frequencies the degree of shielding will depend upon the permeability of the iron, and, as this varies greatly with different kinds of iron and the amount of magnetomotive force used, the results of Figure 34 can be used in a qualitative way only.

At low frequencies where the eddy currents do not assure very effective shielding, the iron may be expected to show up as the superior shield because of the effect referred to in connection with Figure 5. At high frequencies the copper, allowing larger eddy currents to flow, might be expected to show up as the better shield, and such is shown to be the case in Figure 35, which

figure shows how iron and copper compare thruout the range of audible frequencies.

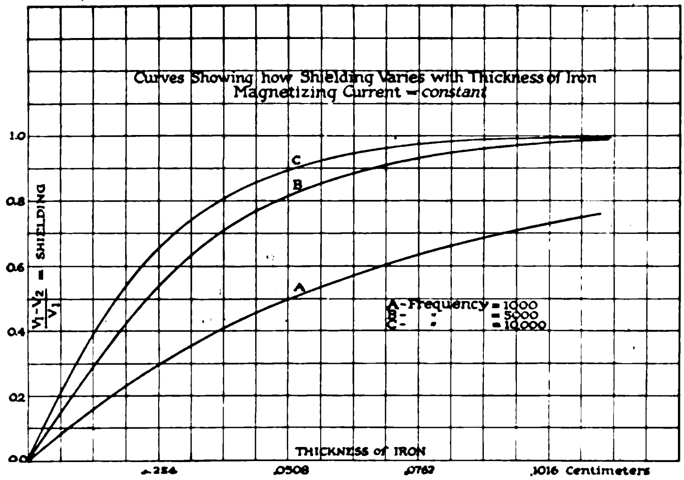


FIGURE 34

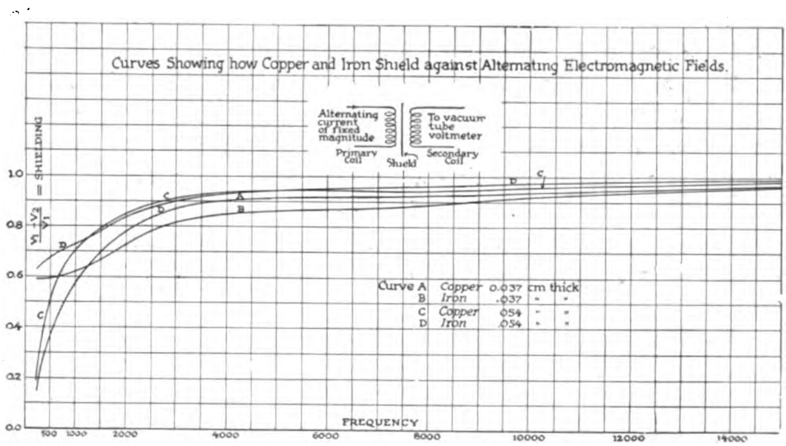


FIGURE 35

CONCLUSIONS

The theoretical and experimental results, when a short-circuited coil was used for a shield, show that:

1. The shielding is zero when either the inductance of the shielding coil or the frequency of the alternating electromagnetic field is zero, and is perfect when the resistance of the shielding coil is zero.

2. When the frequency of the alternating electromagnetic

field and the inductance of the shielding coil remain constant and the resistance in the circuit with this coil is varied, the shielding decreases, approaching zero as the resistance becomes larger.

3. When the resistance and inductance of the shielding coil remain constant and the frequency of the electromagnetic field is varied, the shielding increases, approaching a limit as the frequency becomes very large. This limit is small for large values of the shielding coil resistance and approaches perfect shielding for small values of this resistance.

4. When the frequency of the alternating electromagnetic field and the resistance of the shielding coil remain constant and the inductance of this coil is varied, the shielding increases approaching a limit as the inductance becomes large. This limit is small for large values of the shielding coil resistance and approaches perfect shielding when this resistance is small.

The results, when sheets of non-permeable metals were used as shields, show that:

1. When the resistivity and thickness of the metal remain constant and the frequency of the alternating electromagnetic field varies, the shielding increases approaching values close to unity as the frequency becomes infinitely large.

2. When the frequency of the alternating electromagnetic field and the resistivity of the metal remain constant and the thickness of the shield is varied, the shielding increases and approaches perfect shielding as the thickness increases.

3. When the frequency of the alternating electromagnetic field and the thickness of the metal remain constant and the resistivity is varied, the shielding decreases with an increase in resistivity.

4. Shielding is always accompanied by an increase in the resistance of the shielded circuit; for a given kind of metal at any specified frequency a certain thickness of shield introduces a maximum resistance into the shielded circuit. The thickness of shield which gives maximum added resistance to the shielded circuit decreases as the frequency increases.

5. The effective shielding area of the sheet of metal is that which cuts the electromagnetic lines of force within the coil.

6. The shielding qualities of a sheet of metal are decreased by slits across the effective shielding area of the metal, which constrain the eddy currents to flow in other than the natural paths.

7. The shielding of copper mesh against electromagnetic induction depends on the connections between the wires of the

mesh so that good connections cause good shielding and poor connections cause poor shielding.

8. The results for a permeable metal show that the shielding may be better or poorer than that afforded by copper, depending upon the frequency of the alternating electromagnetic field, the thickness, and resistivity of the metal of the shield.

SUMMARY: An experimental investigation of the shielding of electric and magnetic fields is reported, for both constant and changing fields.

The effect of using iron shells, or sheets, for shielding against the fields of permanent magnets, as well as those set up by electric currents, is considered; the best form for the iron sheets is deduced and an expression for a measure of the shielding action is suggested.

The reason for the leakage of magnetic and electric fields is shown to be due to differences of magnetic or electric potentials in the circuit in which the fluxes are being set up; several cases are cited in which no external fields are set up, as the circuits exhibit no differences in potential.

An expression for the shielding effect of a short-circuited coil is deduced and experimental verification is offered for frequencies between 10^2 and 10^6 cycles per second.

Finally the shielding effect of metal sheets against changing magnetic fields is analyzed and experimental results are given to show how the action depends upon the characteristics of the material of which the shielding plate is made, its thickness, and upon the frequency being used. The effect of slits in the metal sheet, and the value of wire mesh, is indicated.

"THE STRAIGHT-LINE FREQUENCY" VARIABLE CONDENSER*

By

Henry C. Forbes

(Colonial Radio Corporation, New York)

The type of rotary variable condenser having its plates so shaped that the graph of the relation between the frequency of the current in the circuit in which it is connected, and the angle of rotation of the condenser, is a straight line, possesses several advantages. Perhaps the first of these is that a direct scale calibration of the condenser in terms of frequency will be of uniform spacing; that is, each division of the scale will represent a certain number of cycles, this number being constant over the range of the condenser.

Under the system used at the present time in this country, there are assigned approximately one hundred communication channels to the broadcast service, extending from a frequency of 550 kc. to 1,500 kc. (546 to 200 meters, respectively), each with a frequency difference of approximately 10 kc. It is then quite convenient so to construct a variable condenser for use as a tuning element in a transmitting or receiving circuit that each division on the scale shall represent ten kilocycles. Since it is customary to use scales with one hundred divisions to a 180° angle of rotation, this arrangement will permit the frequency range of the broadcast service to be included within the one hundred divisions on the scale and allow a slight over-lap at either end.

The shape of plate for a condenser which will tune a circuit in this manner is dependent upon the constants of the particular circuit in which it is to be used. This will be shown in the development which follows:

The fundamental condition for a straight-line frequency condenser is that:

$$\frac{\Delta f}{\Delta \theta} = -K \tag{1}$$

* Received by the Editor, January 21, 1925.

where Δf = cycles per division
$\Delta \theta$ = angular measure of one division
K = a constant

Expressed as differentials, we have:

$$df = -K\,d\theta \tag{2}$$

In a tuned circuit,

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{D}{\sqrt{C}}$$

where f = frequency
C = capacity of tuned circuit
L = inductance of tuned circuit
$D = \frac{1}{2\pi\sqrt{L}}$, a constant of the circuit for variable C.

Integrating
$$\frac{D}{\sqrt{C}} = -K\theta + B$$

Determining the value of B:
when $\theta = 0$, $C = C_o$

Hence
$$B = \frac{D}{\sqrt{C_o}}$$

and
$$C = \left[\frac{D}{\frac{D}{\sqrt{C_o}} - K\theta}\right]^2 \tag{3}$$

where C_o = capacity in tuned circuit at zero setting of condenser.

The capacity of a multi-plate air condenser neglecting the edge effects is:

$$C = nkA$$

where n = number of dielectric spaces
$k = \frac{10^{-11}}{36\pi d}$, a constant
d = gap between plates
A = active area of each plate.

In the rotary plate condenser this becomes:

$$C = \frac{nk}{2}\int r^2\,d\theta \tag{4}$$

where r = radius vector to edge of rotary plate at the angular setting θ.

Equating (3) and (4), and differentiating:

$$\frac{dC}{d\theta} = \frac{nk}{2}r^2\,d\theta = 2\left[\frac{D}{\frac{D}{\sqrt{C_o}} - K\theta}\right]\left[\frac{DK}{\left(\frac{D}{\sqrt{C_o}} - K\theta\right)^2}\right]d\theta$$

or

$$r=\sqrt{\frac{4\,D^2}{n\,k\,K^2\left[\dfrac{D}{K\sqrt{C_o}}-\theta\right]^3}}$$

To take account of the customary cut-out of the stationary plates, we have:

$$r=\sqrt{\frac{4\,D^2}{n\,k\,K^2\left[\dfrac{D}{K\sqrt{C_o}}-\theta\right]^3}+r_1^2} \tag{5}$$

which is the equation of the desired rotary plate when

$D=\dfrac{1}{2\,\pi\sqrt{L}}$, a circuit constant for variable C.

$K=-\dfrac{d\,f}{d\,\theta}=\dfrac{f_o-f_{100}}{\pi}$, cycles per radian on condenser scale.

n = number of dielectric spaces in condenser.

$k=\dfrac{10^{-11}}{36\,\pi\,d}$, a constant for the capacity of the condenser.

C_o = total capacity in circuit when $\theta=0$.

θ = angle of rotation of rotor plates.

r_1 = radius of cut-out on stationary plates.

d = air-gap between plates.

All dimensions are in centimeters, the inductance is in henrys, the capacity in farads, and the angles are expressed in radians.

The capacity of the condenser alone, at any angular position of the rotor, θ', is:

$$C_\theta=\frac{D^2}{\left[\dfrac{D}{\sqrt{C_o}}-K\,\theta'\right]^2}+\frac{n\,k}{2}\,r_1^2\,\theta'-C_o \tag{6}$$

SUMMARY: The equation for the shape of the rotary plates in a rotary variable condenser is developed so that the frequency—angular setting characteristic is a straight line. The equation for the capacity of this condenser at any angular setting is also given.

CALCULATION OF THE MUTUAL INDUCTANCE OF CO-AXIAL CYLINDRICAL COILS OF SMALL RADIAL DEPTH*

By

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Many radio circuits utilize the mutual inductance between co-axial cylindrical coils of small radial depth.

If a set of curves, similar to Figure 1a, showing the mutual inductance between co-axial circles is available, the mutual inductance of such coils may be estimated quite rapidly by a simple process of summation. The method of constructing such curves is described in a recent publication of the Bureau of Standards[1] and several examples are given.

Figure 1a differs from these curves only in that it is made slightly more convenient for the purpose at hand, by adding intermediate curves. The process of summation referred to is accomplished by dividing the coils into a number of parts, determining the mutual inductance between the parts separately, and then adding the values so obtained to get the total. This is very rapidly accomplished if a templet representing the coils drawn to scale is constructed. Cutting out this templet and sliding it over the curves gives the mutual inductances of the parts and all that remains to be done is to sum them up and make allowance for the scale and numbers of turns in the coils.

Figure 1b illustrates such a templet representing two coils "A" and "B." Coil "A" has a radius of 10 cm. (3.94 in.) and length of 8 cm. (3.15 in.) and contains 85 turns. Coil "B" has a radius of 13 cm. (5.12 in.) and length of 5 cm. (1.97 in.) and contains 62 turns. Coil "A" is drawn to such a scale that its radius corresponds to 1 cm. in Figure 1a. The coil "A" has been divided into four sections, designated a_1, a_2, a_3, etc., and coil "B" has been divided into three sections designated b_1, b_2, etc.

*Received by the Editor, December 24, 1924.

[1] "Formulas, Tables and Curves for Computing the Mutual Inductance of Two Co-axial Circles," by Harvey L. Curtis and C. Matilda Sparks, "Scientific Paper 492, Bureau of Standards."

Let N_A = number of turns in "A"

N_B = number of turns in "B"

n_A = number of arbitrary sections in "A"

n_B = number of arbitrary sections in "B"

r_A = radius of coil "A"

Then the number of turns per section is $\frac{N_A}{n_A}$ for "A" and $\frac{N_B}{n_B}$ for "B" and the mutual inductance between any two sections is given by

$$M_{ab} = r_A \frac{N_A N_B}{n_A n_B} M'_{ab}$$

where M'_{ab} is the mutual inductance of two co-axial circles obtained by application of the templet to Figure 1a. The mutual inductance of the two coils is therefore

$$M = r_A \frac{N_A N_B}{n_A n_B} \Sigma M'_{ab}$$

For the case under consideration

$r_A = 10$	$N_B = 62$	$n_B = 3$
$N_A = 85$	$n_A = 4$	

Finding the values of $M'ab$ for the different sections gives—

$M'a_1b_1 = 5.25$	$M'a_3b_1 = 8.80$
$M'a_1b_2 = 4.25$	$M'a_3b_2 = 6.85$
$M'a_1b_3 = 3.45$	$M'a_3b_3 = 5.50$
$M'a_2b_1 = 6.70$	$M'a_4b_1 = 11.40$
$M'a_2b_2 = 5.40$	$M'a_4b_2 = 8.90$
$M'a_2b_3 = 4.33$	$M'a_4b_3 = 7.00$

and the sum of these is 77.83. Hence the mutual inductance is given by

$$M = 10 \times \frac{85 \times 62}{4 \times 3} \times 77.83 = 341{,}500 \text{ milli-microhenrys.}$$

The accuracy of the results obtained is dependent on the number of sections into which the coils are divided. In general the sections should be taken small enough so that the value of M'_{ab} for the midpoints of the sections will be a good average for that function over the whole of both sections involved.

SUMMARY: This note shows how the mutual inductance of coaxial cylindrical coils of small radial depth may readily be obtained by the use of curves of a type recently described by the United States Bureau of Standards. These curves cover the mutual inductance between coaxial circles, and by a very simple process of summation their usefulness can be extended to coils of small radial depth.

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

ISSUED MAY 5, 1925–JUNE 30, 1925

BY

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1,529,416—F. H. Hendey and R. B. Everson, filed August 3, 1922, issued May 26, 1925.

RADIO DETECTOR AND PROCESS FOR MAKING THE SAME, wherein a detector element of a content of lead sulphide over 90 percent by weight and a content of free lead of over 5 percent by weight is provided.

1,536,453—G. W. Pickard and J. A. Proctor, filed November 8, 1924, issued May 5, 1925. Assigned to Wireless Specialty Apparatus Company, Boston.

VERNIER FOR TUNING REACTANCES more particularly for variable condenser in which an auxiliary condenser plate is rotatably mounted on the shaft which carries the main condenser plates. the auxiliary condenser plate being arranged to co-operate with an independent plate which may be varied in its angular position with respect to the auxiliary plate.

1,536,855—W. G. Houskeeper, filed July 20, 1920, issued May 5, 1925. Assigned to Western Electric Company, Incorporated.

ELECTRON DISCHARGE DEVICE of high power construction, in which the tube electrodes are supported intermediate of the tube by means of a crimped metallic collar surrounding the neck of the glass plant within the tube. The crimped collar forms a substantial mounting for the electrodes within the tube.

1,536,860—P. G. Jacobson, filed July 5, 1922, issued May 5, 1925.

CONDENSER of variable capacity in which the plates are

*Received by the Editor, July 11, 1925.

mounted in arcuate slots distributed along supporting sleeves. The plates enter the slots and are secured therein in spaced relationship.

1,536,954—B. R. Webster, filed April 28, 1924, issued May 5, 1925. Assigned to Reliance Die & Stamping Company, Chicago.

ELECTRIC CONDENSER of variable capacity in which the stationary plates and movable plates are supported by means of rods which extend between the end frame plates forming the condenser. The plates are grooved to fit around the rods for supporting the plates in spaced relationship.

1,536,974—G. A. Rosenfelder, filed March 3, 1924, issued May 5, 1925.

CRYSTAL DETECTOR of cartridge construction in which an insulated barrel is provided with a crystal holder at one end and a resilient mounting for a cat whisker at the other end.

1,536,997—A. E. Wyatt, filed March 13, 1924, issued May 5, 1925.

LOOP ANTENNA FRAME CONSTRUCTION which may be folded into a small space and the supporting arms unfolded and automatically locked in outstanding position for supporting the loop antenna.

1,537,021—H. C. Rentschler, filed November 9, 1918, issued May 5, 1925. Assigned to Westinghouse Lamp Company.

OSCILLATION GENERATOR OF THE ARC TYPE, where the arc takes place between electrodes disposed in an atmosphere of a mixture of hydrogen and mercury vapor at a relatively high pressure for the securing of a fine stream discharge.

1,537,124—W. Lytton, filed February 13, 1923, issued May 12, 1925. Assigned to Lytton, Incorporated, Chicago.

RADIO RECEIVING DEVICE comprising a socket which is arranged to receive either the usual vacuum tube or a base supporting a crystal detector. The connections within the socket are arranged so that a crystal detector carried by the usual electron tube base may be inserted within the socket to make connections with the radio receiving circuits.

1,537,228—J. O. Gargan, filed June 3, 1922, issued May 12, 1925. Assigned to Western Electric Company, Incorporated.

MEANS FOR COOLING CARRIER WAVE APPARATUS, in which the apparatus is mounted in a supporting frame in such manner that air currents may readily move upwardly thru the parts of the set for cooling the apparatus under conditions of operation.

1,537,386—E. M. Tingley, filed November 24, 1920, issued May 12, 1925. Assigned to General Electric Company, Schenectady, New York.

ELECTROSTATIC CONDENSER in which the condenser stack is secured under pressure by clamping members which are directed at right angles to each other and which are secured across opposite sides of the stack. The clamps for the stack also serve as terminals for the condenser.

1,537,660—W. Dubilier, filed January 31, 1924, issued May 12, 1925. Assigned to Dubilier Condenser and Radio Corporation.

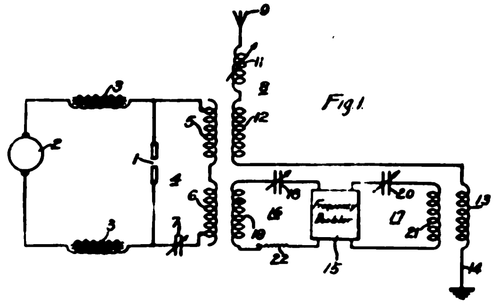
ELECTRICAL CONDENSER in which the stack is secured under pressure by the conjoint action of a flattened tubular casing and tubular rivets which extend thru the stack. The capacity of the condenser may be fixed to a substantially permanent value by reason of the pressure under which the stack is placed.

1,537,561—F. E. Stern and J. C. Randall, filed August 16, 1922, issued May 12, 1925. Assigned to Stern & Company, Hartford.

VARIABLE CONDENSER in which the semi-circular stator plates are supported in rail members positioned between the condenser plates and then the end of the rotatable spindle secured in position with a flat spring member which bears directly against the end thereof.

1,537,609—J. V. L. Hogan, filed May 19, 1922, issued May 12, 1925. Assigned to Westinghouse Electric and Manufacturing Company.

ARC TRANSMISSION SYSTEM, wherein the purpose is to increase the radiation of the system. A non-sinusoidal generator is employed and the frequency of the main wave multiplied and the multiplied frequency fed in phase with a harmonic of the output of the generator to the antenna circuit. In this way the full power of the several frequencies is radiated.



NUMBER 1,537,609—Arc Transmission System

1,537,708—W. Schottky, filed August 27, 1919, issued May 12, 1925. Assigned to Siemens & Halske, Siemenstadt, Berlin.

THERMIONIC VACUUM TUBE in which the reaction between the electron discharge and external magnetic fields surrounding the tube in an amplifier circuit is diminished by arranging in front of the anode an open work or grid shaped conductor insulated from the anode as well as from the controlling grid electrode, and possessing a constant voltage relatively to the cathode. The peculiar action of this protective net is explained by the fact that it protects electrostatically the field in the vicinity of the auxiliary or grid electrode against the field of the anode, while at the same time the field directed from the grid towards the protective net and partly penetrating into the space between cathode and grid causes a considerable part of the electrons to pass thru the grid even in the case where the grid is given a potential lower than that of the cathode.

1,537,856—F. Michels and A. Erisman, filed September 16, 1922, issued May 12, 1925.

CRYSTAL DETECTOR having a tubular container enclosing a crystal holder and a crystal detecting element which is secured in fixed position with respect to the crystal for maintaining the adjustment of the detector under conditions of vibration.

1,537,990—L. Espenschied, filed April 11, 1924, issued May 19, 1925. Assigned to American Telephone and Telegraph Company.

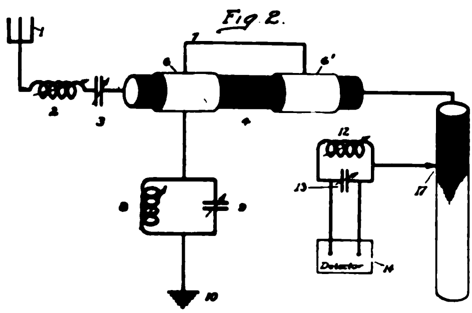
METHOD OF IMPROVING BROADCAST RECEPTION by reducing "fading" effects. It is proposed to reduce the effects of attenuation due to the distance of the point from which the signal is broadcast and also to reduce the fading effect, by producing at

a local station in the area in which is included a large number of receiving points, a carrier wave of exactly the same frequency as that employed by a distant broadcasting station, and then broadcasting the wave to the various receiving sets in the local area. The amplitude of this wave may be made so great that the locally produced carrier will be quite large at each receiving point as compared with the amplitude of the carrier actually received from the broadcasting station. The received signal will then be proportional to the product of the received side band and the carrier transmitted from a point in the neighborhood of the receiving station. This will result in receiving a signal of greater strength than would be possible where the signal depended upon the transmission of both carrier and side band from the distant transmitting station.

1,538,344—E. S. Miller, filed February 1, 1923, issued May 19, 1925.

VARIABLE CONDENSER having fixed and movable plates where the movable plates are supported on a shaft which is carried in a single bearing, the bearing consisting of a pair of opposed frustro-conical portions in which the rotatable shaft may be journaled and adjusted to bring the removable plates intermediate the stator plates.

1,538,466—Louis Cohen and J. O. Mauborgne, filed October 25, 1920, issued May 19, 1925.

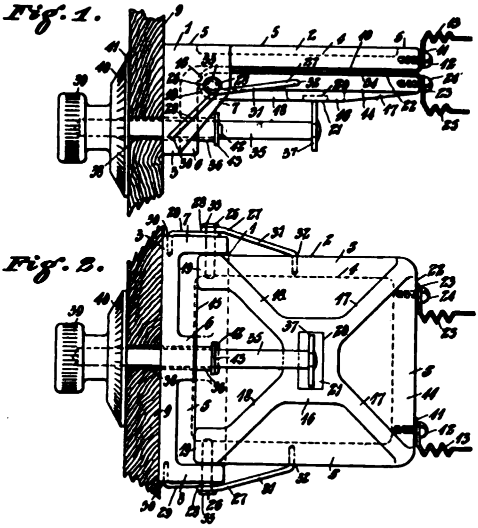


NUMBER 1,538,466—Electrical Signaling

ELECTRICAL SIGNALING for the reception of signals without interference arising from static. The received energy is caused to act upon an ungrounded antenna system and pass thru a wave coil for producing a wave development on the wave coil. The wave coil is operatively associated with adjustable metal tubes which

are changed in position along the wave coil for the best signal reception.

1,538,472—P. Crosley, Jr., filed May 23, 1921, issued May 19, 1925.



NUMBER 1,538,472—Condenser

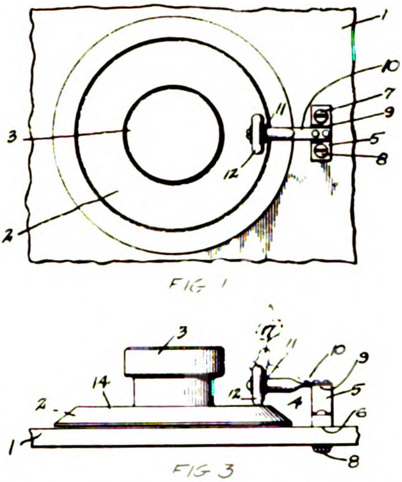
CONDENSER of the book type in which one plate is hingedly mounted with respect to another plate and moved about a pivot with respect to said plate by means of a cam actuated by a shaft member extending thru an instrument panel.

1,538,487—O. Meirowsky, filed April 29, 1922, issued May 19, 1925.

ELECTRICAL CONDENSER of variable capacity in which the rotatable plate is formed by a helical member which may be screwed axially between a fixed plate also formed of a helical member for adjusting the capacity of the condenser.

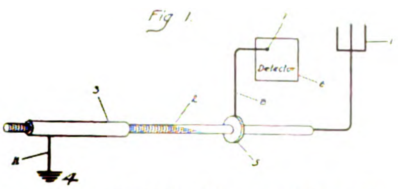
1,538,523—Le Roy W. Staunton, filed June 9, 1924, issued May 19, 1925. Assigned to C. Brandes, Incorporated.

GEARING FOR INSTRUMENT DIALS, in which a small attachable supporting member may be placed upon an instrument panel carrying a resilient member thereon with a rotatable friction wheel at the end thereof arranged to engage the instrument dial for rotating the instrument dial thru small angular distances.



NUMBER 1,538,523—Gearing for Instrument Dials

1538,570—L. Cohen and J. O. Mauborgne, filed June 16, 1920, issued May 19, 1925.



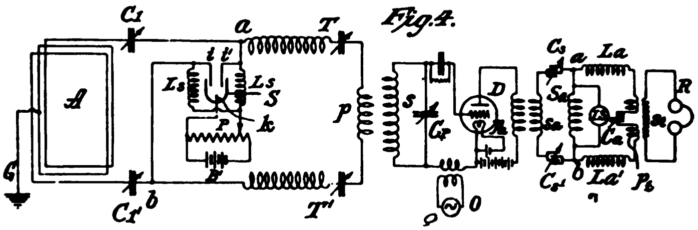
NUMBER 1,538,570—Electrical Signaling

ELECTRICAL SIGNALING system for receiving radio signals, where an antenna is connected to one end of a resonance wave coil with an adjustable grounded metal tube operatively associated with the wave coil. A secondary take-off circuit is coupled with the wave coil for operation of receiving apparatus.

1,538,666—C. M. Small, filed July 8, 1924, issued May 19, 1925.

ELECTRIC SWITCH, which includes a lightning arrestor for the connection of radio apparatus with an antenna system. The lightning arrestor and switch are mounted together and serve to protect the apparatus from excessive currents arising from atmospheric disturbances.

1,538,975—F. K. Vreeland, filed August 6, 1919, issued May 19, 1925.



NUMBER 1,538,975—Stray Elimination in Radio Receivers

STRAY ELIMINATION IN RADIO RECEIVERS, in which a baffle circuit tuned to the signal frequency is provided with a by-pass circuit including a reactance and an intensity selector co-operating to divert preferentially stray energy from the receiving system. The receiving apparatus is connected with the baffle circuit.

1,539,150—O. Von Bronk, filed September 3, 1921, issued May 26, 1925. Assigned to Gesellschaft für drahtlose Telegraphie m. b. H., Hallesches, Berlin.

RECEIVING DETECTOR FOR RADIO TELEGRAPHY, where a generator is provided at the receiving station with the receiving circuit connected to the field of the generator. A circuit is provided connected to the armature of the generator and arranged to resonate simultaneously the frequencies equal to the sum and difference of the signal frequency and natural generator frequency. An audible frequency is thus produced of desired tone quality for observing the transmitted signals.

1,539,402—H. W. Nichols, filed February 24, 1919, issued May 26, 1925. Assigned to Western Electric Company, Incorporated.

MEANS FOR PRODUCING ELECTRICAL OSCILLATIONS, comprising an amplifier having input and output circuits, the circuits of which combined with the amplifier constitute a network for determining the frequency of the generated oscillations independent of any electrical capacity.

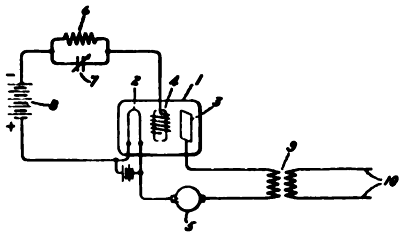
1,540,176—F. P. March, filed June 23, 1922, issued June 2, 1925.

CONDENSER of the variable capacity in which movable plates are mounted upon a sleeve member which is journaled to rotate in end plates which form the condenser frame. The sleeve member extends thru the end plates and the rotatable control knob is secured upon the sleeve member.

1,540,355—R. C. Mathes, filed October 20, 1919, issued June 2, 1925. Assigned to Western Electric Company, Incorporated.

VACUUM TUBE TESTING DEVICE, in which a vacuum tube is mounted upon a vibrating table and subjected to vibratory tests to determine the mechanical rigidity of the mounting of the electrodes with the tube. Circuits are provided for testing the tubes after periods of vibrations to determine the effects of the vibratory tests thereon.

1,540,578—H. C. Thompson, filed July 1, 1921, issued June 2, 1925. Assigned to General Electric Company, Schenectady.



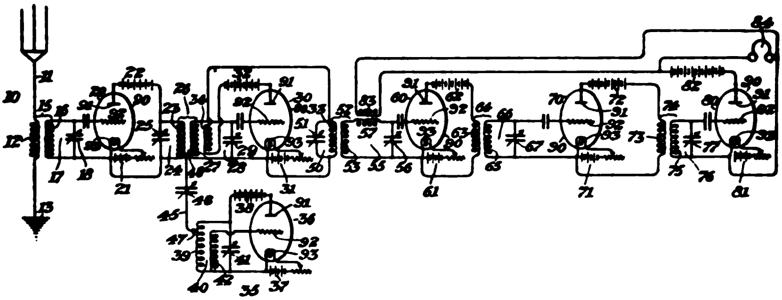
NUMBER 1,540,578—Means for Producing Oscillations

MEANS FOR PRODUCING OSCILLATIONS, wherein a three-electrode electron tube device has an oscillatory circuit connected between the cathode and the grid with a source of potential in the output plate circuit of such value as to permit of the effective emission of impact electrons from the grid in sufficient amount to produce a negative resistance characteristic in the circuit connected thereto over a range of positive potential of the grid. The patent covers the particular values of grid and plate potentials which are desirable for the production of oscillations.

1,540,712—L. Pungs, filed August 23, 1921, issued June 2, 1925. Assigned to C. Lorenz Aktiengesellschaft, Berlin-Tempelhof, Germany.

SIGNAL SENDING APPARATUS IN WAVE TELEGRAPHY, wherein a radio frequency machine is employed having a plurality of syntonizing circuits of different frequency with a coil having an iron core electrically connected with said circuits whereby the coil may be subjected to superposed magnetic continuous current induction, thereby varying the inductivity of the coil and changing the conditions of resonance of the syntonizing circuits.

1,540,881—J. H. Hammond, Jr., filed July 9, 1918, issued June 9, 1925.



NUMBER 1,540,881—Receiving System for Radiant Energy

RECEIVING SYSTEM FOR RADIANT ENERGY, in which a circuit is provided for producing beats for actuating a detector. A circuit is arranged to be controlled by a detector which circuit reacts upon the first circuit for accentuating the beats in the receiving circuit.

1,540,998—H. Plauson, filed January 13, 1921, issued June 9, 1925.

CONVERSION OF ATMOSPHERIC ELECTRIC ENERGY by the use of collecting aerial networks which are elevated in the air and interconnected with a circuit by which static electricity is conveyed to earth in the form of direct current of high voltage and low current strength and converted into electrodynamic energy in the form of high frequency vibrations. The energy derived from atmospheric discharge is intended to be usefully employed for lighting, production of heat, and for use in electro-chemistry.

1,541,566—A. W. Hull, filed May 18, 1921, issued June 9, 1925. Assigned to General Electric Company, Schenectady. New York.

SYSTEM FOR PRODUCING OSCILLATIONS by selecting the potential of the grid and plate electrodes of an electron tube. An emission of secondary electrons from the grid is produced sufficient to cause a negative resistance characteristic in the grid circuit over a range of positive grid potential and thereby causing the generation of oscillations in the grid circuit independently of any coupling between the circuits and a work circuit associated with said plate circuit.

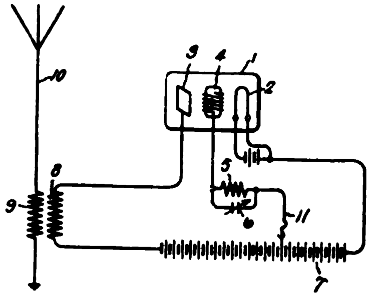
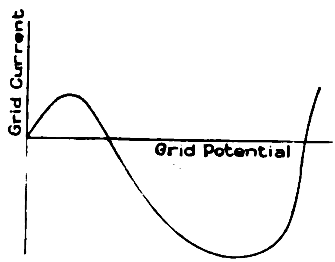
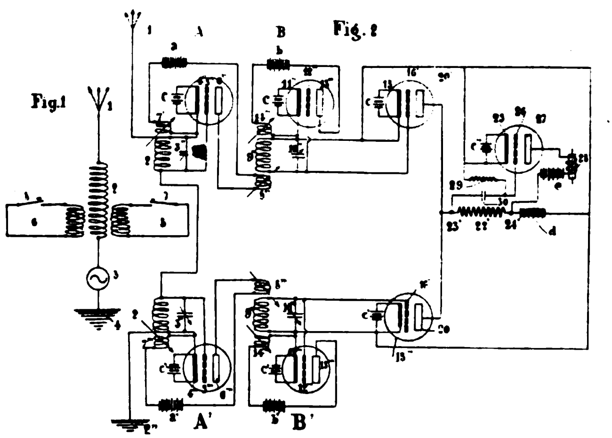


Fig. 2.



NUMBER 1,541,566—System for Producing Oscillations

1,541,608—H. Abraham, filed September 3, 1921, issued June 9, 1925.



NUMBER 1,541,608—Radio Telegraphy

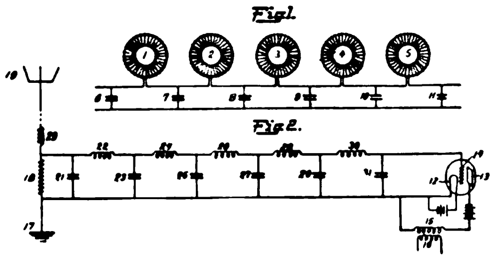
RADIO TELEGRAPHY system, by which several messages can be radiated simultaneously from the same antenna system. The

several transmitters are each connected with the antenna system in such manner that signals are transmitted at separated frequencies without reaction of one frequency upon the others.

1,541,630—W. Dubilier, filed April 22, 1924, issued June 9, 1925. Assigned to Dubilier Condenser and Radio Corporation.

ELECTRICAL CONDENSER FOR IGNITION CIRCUITS, having top and bottom plates of insulation material and separated side plates joining the top and bottom plates and providing terminals. The stack is capable of insertion as a unit between the side plates for properly securing the stack in position.

1,541,845—M. I. Pupin, filed December 11, 1915, issued June 16, 1925. Assigned to Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania.



NUMBER 1,541,845—Electrical Wave Receiving System

ELECTRICAL WAVE RECEIVING SYSTEM, in which the antenna circuit is connected with the receiving circuit thru a recurrent network of similar sections. The network contains damping resistances in each section. The object of the invention is to exclude from the receiving circuit all waves which are not intended to be received.

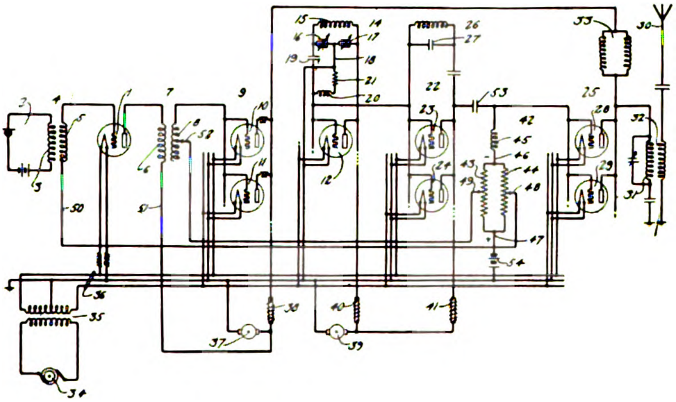
1,542,258—J. McMenamin, filed October 17, 1922, issued June 16, 1925.

MINERAL DETECTOR AND RECTIFIER FOR RADIO RECEPTION, which includes a mass of rectifying material and a plurality of insulated wires grouped at one end to form a compact mass with a flush face and presenting a plurality of fixed contacts opposite to the mineral. The wires are independently brought out for connection in the receiving circuit whereby the most sensitive junction may be utilized in the receiving circuit.

1,542,366—W. R. Brough, filed December 2, 1920, issued June 16, 1925. Assigned to Western Electric Company.

VACUUM TUBE BASE, which consists of a sheet metal shell which supports a thin disc of insulating material in one end thereof the disc carrying terminal pins for the electron tube electrodes which terminal pins are arranged in the corners of a quadrilateral, with two of said terminal pins at different distances from the shell.

1,542,381—J. C. Gabriel and Arvid G. Landeen, filed June 13, 1924, issued June 16, 1925. Assigned to Western Electric Company.



NUMBER 1,542,381—Discharge Device System

DISCHARGE DEVICE SYSTEM for radio transmission, wherein the power amplifier tubes of the system are provided with a potentiometer circuit in shunt with the input circuit of the power amplifier, insuring a relatively large continuous flow of current therethru, with taps taken from a plurality of points on the potentiometer to connections with the grid circuits of a plurality of tubes in the transmitting system for controlling the negative grid potentials of said tubes from a single location in the system.

1,542,385—J. E. Harris, filed October 17, 1920, issued June 16, 1925. Assigned to Western Electric Company.

THERMIONIC CATHODE AND METHOD OF MAKING THE SAME, where the cathode is formed by an alloy containing approximately 95 percent platinum and 5 percent nickel.

1,542,386—R. V. L. Hartley, filed December 14, 1920, issued June 16, 1925. Assigned to Western Electric Company.

VACUUM TUBE DESIGN, in which the characteristic of the vacuum tube is modified so that the space current, when the grid potential is given increasing negative values, approaches a zero value more rapidly than heretofore, thereby giving a more accurate determination of the values of the negative grid potential required to produce zero space current. The arrangement of the tube circuit permits the device to be used in the measurement of alternating current potentials by means of a meter disposed in the output circuit.

1,542,389—W. G. Houskeeper and Wm. R. Brough, filed November 19, 1920, issued June 16, 1925. Assigned to Western Electric Company.

VACUUM INSULATED TERMINAL for an electron tube discharge device where a tubular member is disposed within the vacuum tube and provides a passageway thru which a lead wire extends from the outside of the tube to one of the interior electrodes of the tube.

1,542,724—O. I. Price, filed May 31, 1922, issued June 16, 1925.

RADIO APPARATUS, in which apparatus is mounted upon an instrument panel by means of screw members which pass thru the panel and are embedded directly in the material which supports the instruments. A variable condenser construction is represented in the patent as supported on the rear of the receiving set panel.

1,542,937—J. H. Hammond, Jr., filed May 8, 1920, issued June 23, 1925.

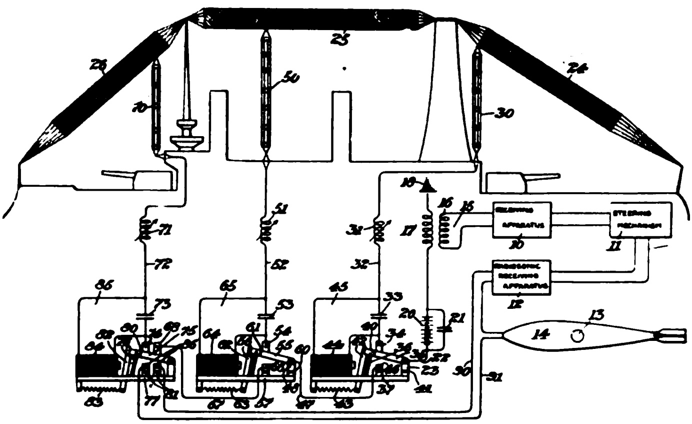
OPTICAL INSTRUMENT for detecting objects in the dark which are invisible to the human eye. A receiver consisting of a thallium oxi-sulphide cell sensitive to light waves is provided and the effect of such light waves upon the object to be detected measured by means of a receiving circuit which discloses the presence of such object.

1,542,938—J. H. Hammond, Jr., filed July 9, 1920, issued June 23, 1925.

AUTOMATIC ANTENNA SYSTEM, particularly adapted for warship use where a plurality of antennas are provided for connection

with a receiving apparatus with an automatic switching system for connecting one antenna with the receiving apparatus in place of an antenna which may be shot away or otherwise destroyed.

Fig. 1



NUMBER 1,542,938—Automatic Antenna System

1,542,995—M. Eastham, filed October 11, 1923, issued June 23, 1925. Assigned to General Radio Company.

CONDENSER of the variable type in which the stator plates are provided with peripheral lugs angularly related to the plane of the plates and arranged to overlap partially correspondingly positioned lugs on one of the stator plates next adjacent thereto. The assembly of the condenser plate spaced by means of the peripheral lugs is intended to reduce manufacturing costs in condenser construction.

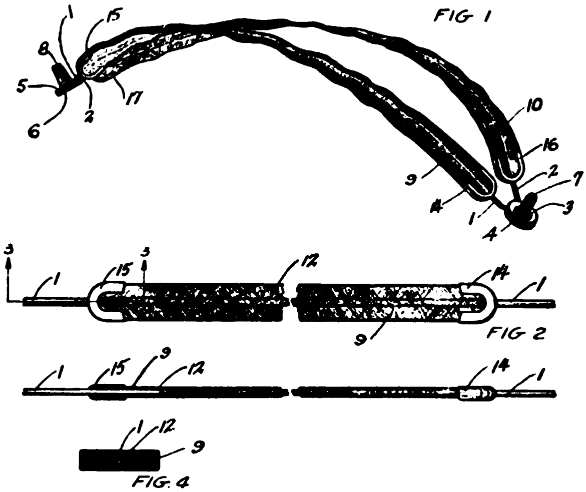
1,543,033—W. O. Snelling, filed May 20, 1922, issued June 23, 1925.

AN ELECTRODE for use in electron tubes in which a grid of electrically conducting material is provided with a light transparent metallic film connecting the individual members of the grid.

1,543,325—F. Dietrich, filed December 8, 1923, issued June 23, 1925. Assigned to C. Brandes, Incorporated.

HEADBAND FOR TELEPHONE HEADSETS, wherein seamless fabric webbing is provided over the wire members which form the headband. The seamless fabric webbing is readily

manufactured and serves as a comfortable construction of headband at the same time that manufacturing costs are considerably reduced.



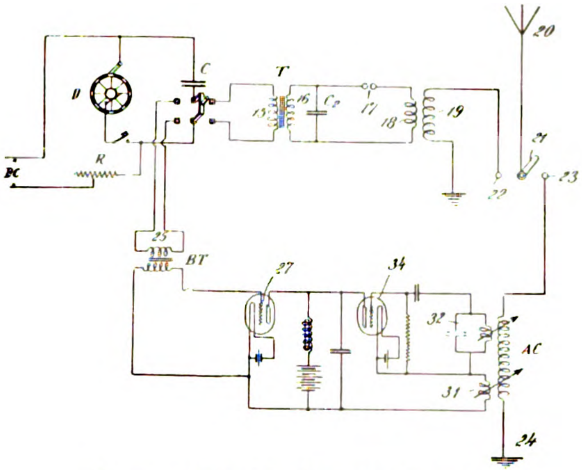
NUMBER 1,543,325—Headband for Telephone Headsets

1,543,326—W. Dubilier, filed February 7, 1921, issued June 23, 1925. Assigned to Dubilier Condenser and Radio Corporation.

CONDENSER AND CLAMP THEREFOR. The condenser comprises a stack of conducting and dielectric sheets with bearing members engaging the opposite faces of the stack. The bearing members are clipped together by substantially circular clamps which fit into grooves in the bearing members.

1,543,475—W. S. Lemmon, filed January 14, 1920, issued June 23, 1925.

RESONANT CONVERTER for producing oscillations for use in radio signaling systems where a direct current source is provided and inductance and capacity elements connected across the terminals of said source with an interrupter in parallel therewith. An oscillatory circuit tuned to the frequency of the alternating current to be produced is connected in series between the source and the inductance and capacity elements and arranged to cooperate with the interrupter for producing oscillations which may be impressed upon the signaling circuit.



NUMBER 1,543,475—Resonant Converter

1,543,726—K. Rottgardt, filed August 26, 1921, issued June 30, 1925. Assigned to Westinghouse Electric and Manufacturing Company.

RADIO TRANSMISSION RECORDING SYSTEM, in which the signals which are transmitted by means of a hand key at the transmitting station are simultaneously recorded by a Morse writer. The Morse writer is controlled by the operation of the hand key to make a record of the signals being transmitted.

1,543,872—S. Ruben, filed April 13, 1923, issued June 30, 1925.

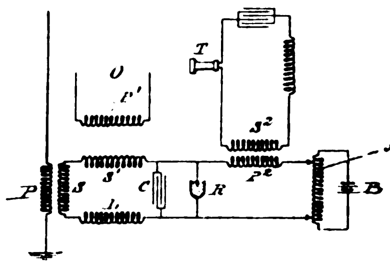
VACUUM TUBE APPARATUS for making and breaking a circuit by relay operation. The discharge device contains a cathode and a plurality of anodes located in the vacous space with a separate external circuit connecting the cathode with either of the anodes. Another circuit is provided within the tube which includes a gap located in said vacuous space and arranged to be closed and opened by the expansion and contraction respectively, of one of said anode elements. The device is used for controlling any desired electrical circuit.

1,544,081—F. K. Vreeland, filed (original) January 2, 1907. This application filed October 27, 1915, issued June 30, 1925. Assigned to Vreeland Apparatus Company.

TRANSMITTING INTELLIGENCE BY RADIANT ENERGY and receiving such signals by exciting a receiving circuit by the received wave energy and also by an auxiliary current of a fre-

quency related to the frequency of the received energy as to produce beats therewith, and then rectifying and using this composite current to operate a receiving instrument, whereby to amplify the effect of the received signal. This patent contains broad claims on heterodyne reception.

Fig. 2



NUMBER 1,544,081—Transmitting Intelligence by Radiant Energy

1,544,102—P. O. Pedersen, filed June 21, 1919, issued June 30, 1925. Assigned to Poulsen Wireless Corporation.

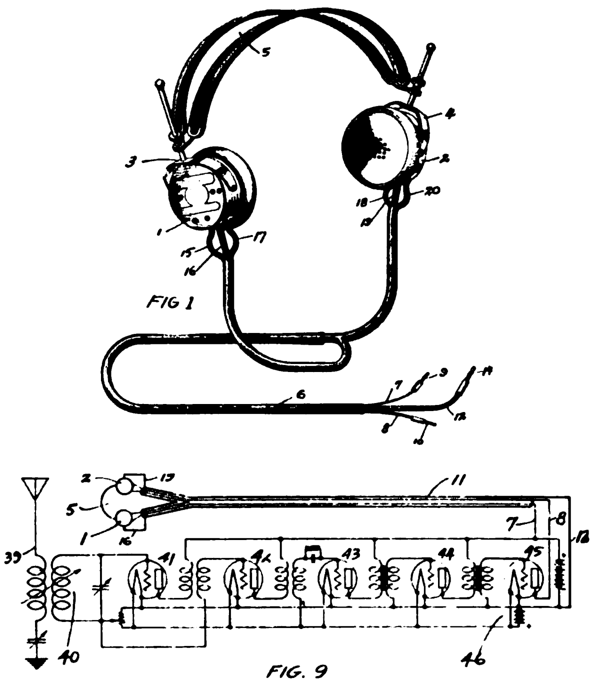
AN ARC GENERATOR, including a pair of electrodes between which the arc is formed with a magnetic blow-out for extinguishing the arc. The shape of the electrodes is such that the arc is caused to increase in length periodically between the electrodes before it is blown out.

1,544,133—C. C. Culver, filed June 29, 1921, issued June 30, 1926.

A TRIDIMENSIONAL RADIOCOMPASS for use in aircraft where a plurality of coils are provided forming a plurality of dihedral angles, the axes of said dihedral angles being angularly disposed with respect to each other and to the longitudinal axis of said aircraft. By use of a multiple number of coils an exact direction may be determined independent of the loading or angular position of the aircraft during flight.

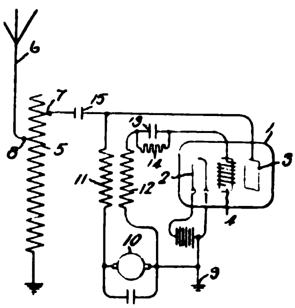
1,544,136—F. Dietrich, filed November 5, 1923, issued June 30, 1925. Assigned to C. Brandes, Incorporated.

ELECTRICAL CONNECTION FOR TELEPHONE HEADSETS, where a flexible metallic shield surrounds the telephone conductors with connections between telephone conductors and the magnet bobbins within the receiver. The flexible shield terminates in a stay cord which is utilized to remove the strain from the telephone conductors which lead into the telephone receiver casing.



NUMBER 1,544,136—Electrical Connection for Telephone Headsets

1,544,202—W. C. White, filed January 20, 1921, issued June 30, 1925. Assigned to General Electric Company.



NUMBER 1,544,202—System for Producing Oscillations

SYSTEM FOR PRODUCING OSCILLATIONS by use of an electron tube which has an oscillating circuit coupled to a shunt circuit between its cathode and anode. The input and output circuits of the tube are coupled, but the frequency of the oscillations is determined principally by the constants of the oscillating circuit.

1,544,486—J. Sedlak, filed May 31, 1924, issued June 30, 1925.

A Switching Device for Radio Sets, in which connections are made to a panel board on the rear of a radio receiving set by means of a multiple blade switch co-operating with contacting clips which connect with the internal circuits of the radio receiver.

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 13 | OCTOBER, 1925 | Number 5

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GENERAL INFORMATION

The Proceedings of the Institute are published every two months and contain the papers and the discussions thereon as presented at the meetings in New York, Washington, Boston, Seattle, San Francisco, or Chicago.

Payment of the annual dues by a member entitles him to one copy of each number of the Proceedings issued during the period of his membership.

Subscriptions to the Proceedings are received from non-members at the rate of \$1.50 per copy or \$9.00 per year. To foreign countries the rates are \$1.60 per copy or \$9.60 per year. A discount of 25 per cent is allowed to libraries and booksellers.

It is understood that the statements and opinions given in the Proceedings are the views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole.

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INSTITUTE ACTIVITIES

Technical Papers Being Prepared for Early Presentation at New York Meetings

Among the important papers now being written for presentation at meetings of the INSTITUTE in New York, during the Fall and Winter months, are the following: "Sources of 'A' and 'B' Power for Broadcast Receivers," by W. E. Holland; "Progress in Short Wave Transmission," by Frank Conrad; " 'B' Battery Sources for Broadcast Receivers," by Harry Houck; a paper on "Primary Battery Sources of Power," by W. B. Schulte; "Radio Engineering Instruction Methods," by C. M. Jansky, Jr.; "Crystal Control in Broadcasting," by M. C. Batsel and G. L. Beers; "Automatic Reception of Time Signals," by L. A. Hazeltine; "Radio Interference Mitigation," by J. O. Smith, and "Testing for Trouble in Radio Broadcast Receivers," by Lee Manley.

Institute Emblems, or Badges

All Fellows, Members and Associates of the INSTITUTE should provide themselves with a badge of membership. The badge is made in the form of a class-pin and is of 14-karat gold, beautifully finished in enamel and gold lettering. The Fellows' badge is of blue lettering on a gold background, Members' badges in gold letters on a blue background, and Associates' badges in gold lettering on a maroon background. When desired, the member's initials are engraved on the reverse side of the badge.

Badges sell for $3.00 each and may be procured from the Secretary, THE INSTITUTE OF RADIO ENGINEERS, 37 West 39th Street, New York.

Papers at Section Meetings

Members of the INSTITUTE are advised that technical papers dealing with radio subjects may be scheduled for presentation at Section meetings, when desirable, instead of being presented at New York meetings of the INSTITUTE. Sections now are established at Washington, D. C., Boston, Massachusetts, Chicago, Illinois, San Francisco, California, and Seattle, Washington.

Papers intended for presentation at Section meetings may be forwarded to the Chairman, Meetings and Papers Committee,

THE INSTITUTE OF RADIO ENGINEERS, 37 West 39th Street, or to the Section Chairman.

M. I. Pupin

Dr. M. I. Pupin, elected President of the American Institute of Electrical Engineers for the year beginning August 1, 1925, is the second engineer to have attained the office of President of the American Institute of Electrical Engineers and President of THE INSTITUTE OF RADIO ENGINEERS.

Dr. Pupin was President of the I. R. E. in the year 1917. The other A. I. E. E. President who served as President of THE INSTITUTE OF RADIO ENGINEERS is Dr. A. E. Kennelly. He was at the head of the A. I. E. E. in 1898-1900, and of the I. R. E. in 1916.

Auxiliary Radio Language

The recently held International Conference of Radio Amateurs, at Paris, France, appointed a committee to study and report upon the problem of an auxiliary language for use in communication, correspondence, and comversation between members of the international units of the organization. About twenty artificial languages were considered, including Esperanto and Ido, as well as a few national languages. After considerable debate, in which the Scandinavian representatives strongly urged the adoption of English as the approved auxiliary language, the committee reported in favor of Esperanto.

Advertisements

Members of the INSTITUTE will have noted the improved make-up and display of the advertising pages of the PROCEEDINGS in recent issues. The value of the PROCEEDINGS as an advertising medium has made it possible for the INSTITUTE to procure copy for a full-page advertisement from each advertiser, in each issue.

Direct benefits will follow if, in writing to advertisers on the subject of apparatus listed or described, members of the INSTITUTE will mention that the advertisement was seen in the PROCEEDINGS.

Digests of United States Radio Patents

For some time past the PROCEEDINGS has carried a "Digest of United States Patents Relating to Radio Telegraphy and Telephony," in the last fifteen to twenty reading pages of each issue. The cost of publishing this matter is considerable and the question will shortly be before us as to whether it is economically desirable to continue the publication of this information.

Members of the INSTITUTE who are interested are asked to

advise the Secretary whether or not this particular information is of real value to them. Decision in the matter will be based upon the views expressed or written to the Secretary.

Institute Technical Papers

Members of the INSTITUTE desiring to present papers should keep in mind that papers intended for presentation at meetings, or for publication, must be forwarded to the Chairman of the Meetings and Papers Committee, THE INSTITUTE OF RADIO ENGINEERS, 37 West 39th Street, New York, at least sixty days before the date of presentation. This is necessary in order properly to provide for editing, scheduling, proof-reading, and printing.

Toronto Section

The first meeting of the proposed Toronto Section of THE INSTITUTE was held at the University of Toronto on the evening of September 25th.

Additions to Membership

At the September meeting of the Board of Direction there were elected to membership in THE INSTITUTE one Fellow, thirty Members, ninety-nine Associates and nine Juniors. Ten of the Members were transferred from the Associate grade.

Meeting of the Board

The September meeting of the Board of Direction was held at INSTITUTE headquarters, 37 West 39th Street, New York, on the evening of September 1st. Those in attendance were: J. H. Dellinger, President; Donald McNicol, Vice-President; Alfred N. Goldsmith, Secretary; W. F. Hubley, Treasurer, and the following Managers: J. V. L. Hogan, A. E. Reoch, Melville Eastham, and A. H. Grebe.

THE EFFECT OF THE SOLAR ECLIPSE OF JANUARY 24, 1925, ON RADIO RECEPTION*

By

Greenleaf W. Pickard

(Consulting Engineer, The Wireless Specialty Apparatus Company, Boston, Massachusetts)

(Communication from the International Union of Scientific Radio Telegraphy)

During the brief interval of totality, a solar eclipse cuts off nearly all radiation over an area of several thousand square kilometers, and for an interval of several hours there is a material reduction of light over a considerable part of the earth's sunlit hemisphere. The actual light intensity within the central shadow is of the same order as full moonlight; a reduction of of about one hundred thousand times, while in the adjacent portions of the penumbra over nine-tenths of the light is cut off. It would naturally be expected that the effect of an eclipse upon radio communication would be in some wise similar to a night effect, altho less in magnitude because of its shorter duration and smaller area.

Altho there have been solar eclipses ever since this earth had a moon, and radio communication for over a quarter century, it is only in the past twelve years that observations of the eclipse effect upon radio reception have been made. I have appended to this paper the rather meagre bibliography of this subject, which altho distinctly contradictory at first reading, nevertheless shows that a solar eclipse has a distinct effect upon the signal strength from a distant station. In the majority of reports of observations made during past eclipses, the signal intensity has increased during the middle of the eclipse, altho in a few instances decreased reception was observed. Most of the prior observations were merely ear estimates of signal intensity, and the large number of negative observations obtained clearly indicates that in general the eclipse effect is small. The ear is grossly insensitive to slight and relatively slow changes in sound intensity; unless the change occurs quite abruptly a two-to-one variation will usually escape notice. Furthermore, the earlier eclipse

*Received by the Editor, March 25, 1925. Presented before The Institute of Radio Engineers, New York, April 1, 1925.

observations were mostly made on spark transmission, and we recognize today that such highly damped radiation does not show the marked fading, sunrise and sunset effects characteristic of continuous wave transmission, and for this reason would be less apt to show a well-defined eclipse effect. The explanation of this difference is a rather obvious one, if we assume these effects are due to interference between different transmission paths to the receiver, and consider the matter from an optical standpoint. If we pass monochromatic light into any form of interferometer, the interference fringes obtained are sharply defined, but if white light is used the fringes become overlapping spectra, and definition is impaired or lost. In conducting radio transmission over considerable distances at night, we appear to be using a gigantic interferometer, with some agency beyond our control continuously and irregularly changing the length of one of the paths to the receiver, and therefore, the penalty we now pay for the use of the otherwise desirable continuous or monochromatic radiation is a marked accentuation of these bothersome effects. For so long as we employ point sources of monochromatic radiation, with maximum radiation in the horizontal plane, and with single point receivers, every broadcasting station will have its zone of severe fading, and radio direction finders will continue to be erratic around sunrise and sunset. It is perhaps fortunate that we have not yet applied the simple remedies for fading to our broadcasting system, for if this had been done, there would have been no eclipse effect for me to investigate.

In making my plans for the radio observation of the eclipse of January 24th, 1925, I therefore decided to confine my attention to continuous wave transmission, and to use continuous recording at as many points as possible. I also decided to center my attention on those transmission frequencies which I had previously found gave the greatest night effect, and so far as was possible to space transmitters and receivers at the distance of maximum fading; that is, to frequencies within the present broadcasting band, and distances of the order of two hundred kilometers. In order that the higher frequencies should also be observed I asked station 2XI at Schenectady to transmit at approximately four megacycles, which was recorded in New York City, and observed at many other points. At the other end of the radio spectrum quantitative measurements by the telephone comparator method were made under Dr. Austin's direction at Washington on the 57-kilocycle radiation from station 2XS at Rocky Point, Long Island, and by the American

Telephone and Telegraph Company on trans-Atlantic reception.

As the object of my work was to record electric field changes, and not directional effects, open antennas were used at all recording stations, loosely coupled to super-heterodyne receivers. The plate circuit of the second detector was opened and the primary of an intermediate frequency transformer inserted, and the secondary w nding of this transformer was closed thru a crystal detector and the recorder galvanometer. The filament and plate voltages were held rigidly constant, and there was no observable change in voltage amplification during the five days of the eclipse schedule. The large open antennas employed also increased the stability of the receivers, for with the relatively large input available a very moderate voltage amplification sufficed. As I did not know how large the eclipse effect might be, I provided a variable coupling from the antenna to the amplifier, calibrated in terms of galvanometer deflection. In this way not only could the deflection of the galvanometer be adjusted either way by a known amount, but the input to the receiver and also the output to the crystal detector was always kept within definite and small limits. The manual type of recorder devised by Mr. H. S. Shaw[1] was used at all of the recording stations, altho at a number of my observing stations frequent galvanometer readings were used in its place.

Many other groups and individuals made valuable radio observations during the eclipse, altho these were in the main of a qualitative character, depending chiefly upon aural estimates of signal intensity. Prominent among these was the "Scientific American," which by an ingeniously timed transmission from four broadcasting stations reached a well-organized group of several thousand broadcast listeners. The preliminary analysis of the "Scientific American" reports, to which I shall later refer, has brought out some very striking facts. The Zenith Radio Company placed a portable broadcasting station at Escanaba, Michigan, almost in the center of the path of totality, while the Edison Light Company of Boston installed their portable station WTAT on the Coast Guard Cutter "Tampa," and broadcast an eclipse program from a point on the Atlantic Ocean just south of Marthas Vineyard, Massachusetts, and near the center of the shadow path. Excellent directional observations were made at Ithaca, New York, under the direction of Prof. Merritt of Cornell University, using stations WGY and WEAF, and a well-

[1] "Short Period Variations in Radio Reception," Pickard, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, Volume 12, April, 1924. Figure 2, page 123.

marked eclipse effect was noted. This paper must, however, deal principally with the records and observations made by my own group of stations, as the mass of outside data—some of which is even now coming to hand—is so vast that many months will be required for any useful analysis aad correlation.

Figure 1 shows that portion of the eclipse network with which this paper deals. At the extreme western end is the Escanaba station operating at 1,120 kilocycles. Aural observation of their transmission by hundreds of broadcast listeners gave as the principal eclipse effect a strengthening of the signal approximately coinciding with maximum shadow. It is unfortunate that I did not have time to arrange for at least one continuous record from Escabana, as the western end of the totality path was traversed by the shadow very shortly after sunrise, and the night effect was probably very strong at that time.

Broadcasting station WGR, of the Federal Telephone and Telegraph Company at Buffalo, New York, operating at a frequency of 940 kilocycles; WGY of the General Electric Company at Schenectady, New York, 790 kilocycles; 2XI, the experimental station of the General Electric Company, also at Schenectady, and transmitting at 4 megacycles; and WBZ of the Westinghouse Electric and Manufacturing Company, at 890 kilocycles, at Springfield, Massachusetts, maintained a constant and lightly modulated transmission from 7.30 to 11.00 A. M., 75° Meridian Time, on January 22nd, 23rd, 24th, 25th, and 26th, 1925. Station WEAF of the American Telephone and Telegraph Company, at lower New York City, operating at 610 kilocycles, with 250-cycle tone modulation, transmitted from 4.00 to 11.00 A. M. on January 23rd, 24th, and 25th.

Under my personal direction at Ithaca, New York, records were made of the transmission from WGR, WGY, and WEAF. At Schenectady, New York, Mr. W. C. Lent, of Union College, made galvanometer readings of reception from WBZ on the eclipse morning. At Hamilton, Massachusetts, Mr. F. W. Dane, of the Wireless Specialty Apparatus Company, made galvanometer readings of reception from WGY on the mornings of January 24th and 25th. At Fitchburg, Massachusetts, under the direction of Mr. T. Parkinson, of the Bureau of Standards, readings and records of reception from WBZ and WGY were taken. At Leominster, Massachusetts, near Fitchburg, Mr. H. Powers took readings from WGY. Massachusetts Institute of Technology station 1XM, at Cambridge, Massachusetts, under the direction of Prof. A. E. Kennelly, took galvanometer readings of reception

from WBZ on January 24th and 25th. Mr. A. F. Murray made records on the eclipse morning of stations WGY, WEAF, and WTAT at Newport, Rhode Island. At Middletown, Connecticut, records were made under the direction of Mr. H. S. Shaw from WGY and WBZ. At Easthampton, Long Island, a phonograph record of reception from WGY was made by Dr. E. E. Free. At the laboratory of the Radio Corporation of America, in upper New York City, and well within the path of totality, records of WGY and 2XI were made by Mr. Arthur Van Dyck, on the mornings of January 22nd, 23rd, 24th, 25th, and 26th, under the direction of Dr. Alfred N. Goldsmith. At Washington, D. C., Dr. C. B. Jolliffe, of the Bureau of Standards, made records of reception from WGY on the 22nd, 24th, 25th, and 26th, while at the U. R. S. I. laboratory in Washington Dr. L. W. Austin supervised the measurement of field intensity from 2XS on the mornings of the 23rd and 24th.

At Philadelphia, Pennsylvania, galvanometer readings were taken under the direction of Prof. Bazzoni of the University of Pennsylvania of reception from WGY and WEAF. Considerably before the time of totality, reception at Philadelphia had fallen to its flat day-time level, and the curves do not show any effect which could be attributed to the eclipse.

At McGill University, Montreal, Canada, Mr. E. L. Bieler observed reception from WEAF. At Montreal, however, the signal fell to the background level after sunrise, and the curve does not show any effect which could be laid to the eclipse.

At Annapolis, Maryland, Mr. G. D. Robinson made galvanometer readings at one minute intervals from WGY, on the mornings of the 24th and 25th. His results show on the eclipse morning a distinct minimum during the middle of the eclipse, which was only 95 percent complete at his point.

At the higher frequencies, in addition to the records made at New York City from 2XI, this same station was observed by Mr. F. W. Dunmore at the Bureau of Standards in Washington, the principal effect noted being a slight decrease in signal during the maximum shadow. Major Mauborgne of the Signal Corps, observing in Washington, 3.5-megacycle signals from Canadian 9AL at Toronto, observed a marked increase in signal during maximum eclipse at Washington. Under the direction of Dr. A. H. Taylor observations were made at Bellevue (near Washington) and elsewhere on two groups of frequencies, one ranging from 3 to 5 megacycles, and the other above 5 megacycles. The former group of frequencies showed an increase of signal strength

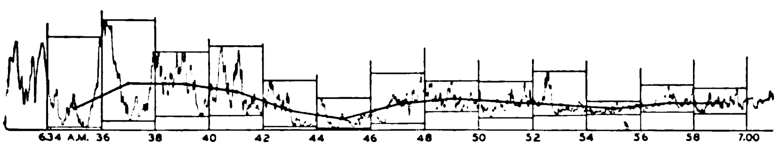
during the eclipse and a decrease as the sunlight returned, while the higher frequencies, particularly those about 7.5 megacycles, were effected in reverse manner, the shadow decreasing and the sunlight increasing the signal strength. Miss E. M. Zandonini of the Bureau of Standards, observing Washington reception of signals from a 2-megacycle station at Newburgh, New York, noted that at the time of maximum shadow in Washington there was a brief sharp increase in audibility. At the present time my analysis of the high frequency data is so far from complete that I can only give this general conclusion: For those frequencies and distances which are better by night than by day, the eclipse improved reception; for frequencies and distances which are normally better by day than by night, the eclipse decreased reception. In other words, the eclipse effect for the higher frequencies appears to be a pure night effect.

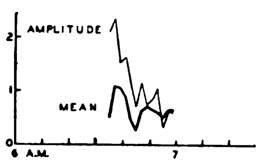
Whenever we record reception from a distant broadcasting station thru the day and night, we find that the record for at least an hour or so before and after noon is quite flat, that is, relatively free from fluctuations. But an hour or more before sunset the fluctuation amplitude begins to increase, and by an hour or two after sunset reaches approximately its normal and high night time value. This continues thru the night, but about an hour before sunrise the fluctuation amplitude begins to fall, as does also, tho to a lesser degree, the mean field, until at about sunrise the mean field has reached nearly its day time level, and the fluctuations have been reduced to a relatively small amplitude. But there are frequent exceptions to this, of which the eclipse morning at most receiving points was one, where the fluctuation amplitude remained high for a long time after sunrise.

We must admit that there is as yet no entirely satisfactory explanation of these phenomena. But most of our transmission hypotheses are in agreement in considering the night field at moderate distances from the transmitter—200 kilometers for example—as made up of two components, one transmitted along a direct or low level path and the other by an indirect or high level route, the second path either changing continually in actual length or in phase. At short distances—five kilometers for example—the energy reaching the receiver by the indirect or high level path is probably negligible, while at great distances—1,000 kilometers for example—the high level path is the principal agent in forming the field at the receiver. According to this general assumption, a record taken thru dawn and sunrise from a station two hundred kilometers distant, and

showing the usual gradual decrease of fluctuation amplitude, may be resolved into two components, one consisting of a curve of mean values, and the other of fluctuation about these mean values. The mean field, taken over any interval which is long as compared with the most prominent fluctuation period, may be interpreted as an approximate measure of the reception along the direct path, while the fluctuation amplitude represents the reception by the indirect or high level path. The decrease of fluctuation amplitude as the sunlight begins to fall thru the air may be taken to mean that either the high level path itself is in some way impaired by the effect of the sunlight, or that it is masked by increasing ionization below, which would attenuate the waves both in their ascent and descent.

WEAF 610 KC AT ITHACA, JAN. 23, 1925.





THIS PORTION OF THE RECORD SHOWS THE MOST MARKED PORTION OF THE CHANGE FROM NIGHT TO DAY RECEPTION. OWING TO THE EXPANDED TIME SCALE OF THE RECORD IT IS NECESSARY TO REPLOT ON A SHORTER SCALE TO SHOW SLOW CHANGES. THE RECORD IS DIVIDED INTO TWO MINUTE SECTIONS AND THE AREA BETWEEN THE CURVE AND THE ZERO LINE MEASURED BY POLAR PLANIMETER, WHICH WHEN DIVIDED BY THE BASE GIVES THE MEAN VALUE FOR THAT SECTION. THIS IS REPEATED FOR EACH SECTION AND HORIZONTAL LINES ARE DRAWN MARKING THE AMPLITUDE OF THE FLUCTUATIONS. AT THE LEFT IS PLOTTED ON A REDUCED TIME SCALE THE MEAN AND FLUCTUATION VALUES TAKEN FROM THE RECORD STRIP ABOVE.

FIGURE 2

I have applied this method of analysis to all of my eclipse records, and the process by which this was done is sufficiently shown in Figure 2. The original record, a portion of which appears in the upper part of this figure, was taken on January 23rd at Ithaca, from station WEAF. For that particular morning this part of the record contains the steepest part of the fall in fluctuation amplitude. As shown in the figure, the record strip is first divided into two-minute sections, and the area between the curve and the base or zero line is measured with a planimeter. This area, divided by the length of the base, gives the mean height of the curve in that section. Repeating this process for each section gives a series of mean values, which when linked give the heavy-line curve shown on the record strip. Also, as indicated on the figure by the short horizontal lines, the amplitude of fluctuation in each section is measured.

The original record, made by the manual recorder, is on a

rather drawn-out time scale, the motion of the paper being approximately 1.6 cm. per minute, so that long period small amplitude changes do not show very well. Therefore, as shown in the lower left-hand part of this figure, the separated mean and amplitude values are replotted on a greatly condensed time scale.

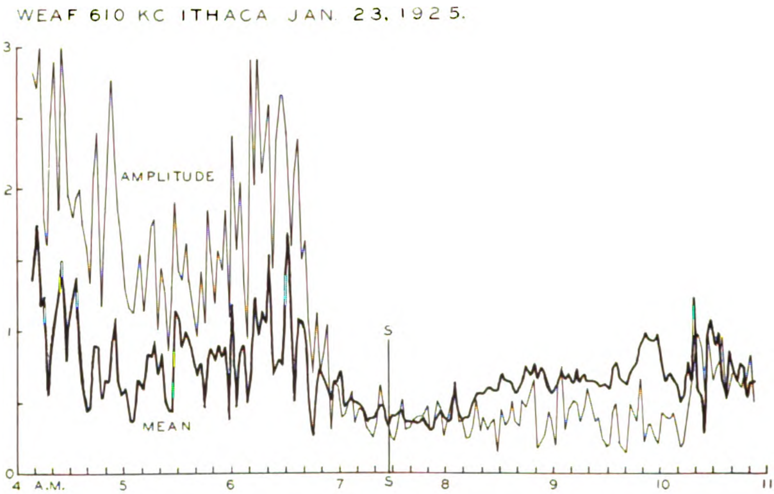


FIGURE 3

The entire morning record, of which Figure 2 is a portion, is replotted in Figure 3. The vertical line S-S marks sunrise at Ithaca, and, so far as the curve is concerned, apparently determined a "low." About an hour before sunrise there is a marked peak in both curves, and preceding this, about two hours before sunrise there is another low. This is quite a normal before-sunrise performance, which was in fact repeated for this station on the eclipse morning.

After sunrise, however, mornings differ greatly. At Ithaca the fluctuations died out soon after sunrise on the 22nd, 23rd, and 25th, while continuing appreciably for over two hours after sunrise on the 24th and 26th. At other recording points the mornings did not arrange themselves in the same order. At Washington the 25th showed the maximum persistence of night effect thru the forenoon, while at New York City all the forenoon records, save the 24th and to a lesser degree the 22nd, are quite flat. If this difference is due to varying ionization below the high level path, it is entirely possible that lower level conditions, even those below the isothermal layer, are involved, in which

event we may find some interesting correlations with weather. Another possible deduction from the frequent persistence of the night effect thru the forenoon is that sunlight does not entirely wreck the high level path, but merely masks this path by a low level ionization, which varies from day to day. In other words, the path may be there all the time, but the waves are prevented by low level conditions from either getting up to it, or coming down again. In the morning a station far to our west, and still in darkness, might deliver its waves quite freely to the high level path, but at our sunlit receiving point they would pass over us unobserved.

WEAF 610 KC AT ITHACA, JAN. 24, 1925.

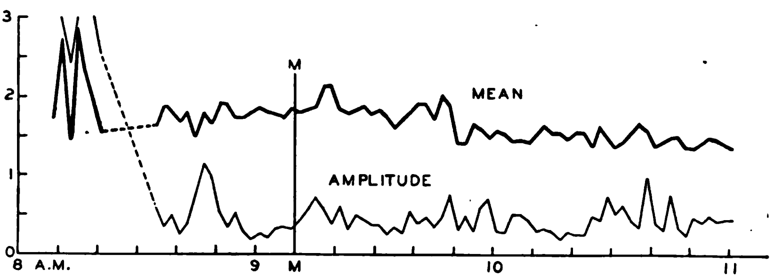


FIGURE 4

Coming now to the eclipse records, Figure 4 is Ithaca reception of WEAF on the eclipse morning. The dotted-line portion of this record between 8.22 and 8.35 A. M. represents a thirteen-minute breakdown on the part of WEAF. The mean curve shows a slight rise, reaching a maximum about ten minutes after the middle of the eclipse, which is indicated on all the eclipse records by a vertical line M-M. The amplitude curve, on the other hand, shows a distinct dip before the middle, and a peak shortly after. It might well be said of this record that similar fluctuations occur at other times, as for example at 10.25 A. M., and so, standing by itself, this record must be considered inconclusive.

Figure 5, of WGY reception at Ithaca, altho quite similar to Figure 4, shows a fall and rise of both mean and amplitude curves which is not duplicated at other points on this record. The effect is most marked, as might be expected, in the amplitude curve, which reaches a minimum value shortly before the middle of the eclipse, and then rises to a maximum some twelve minutes after.

The record of WGR reception at Ithaca, shown in Figure 6, is a very striking one, the principal feature being a high and

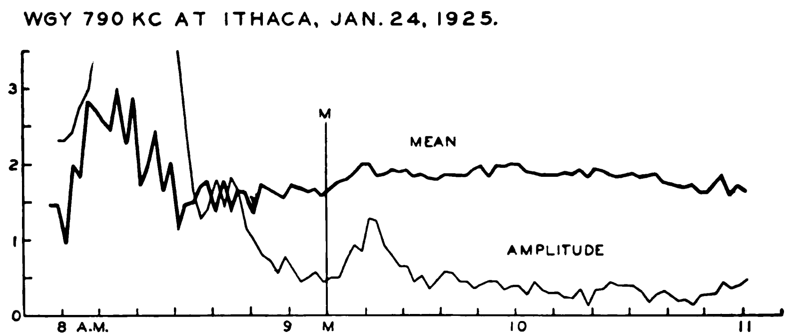


FIGURE 5

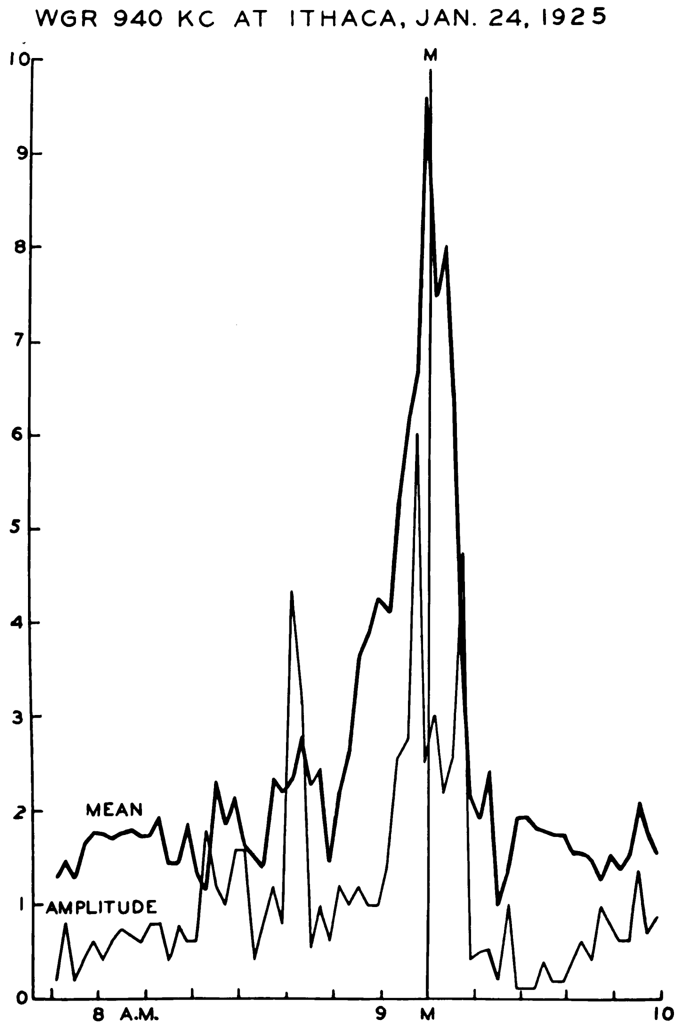


FIGURE 6

sharp peak practically at the middle of the eclipse. This is the maximum effect obtained at any of my recording stations, and at first sight is quite different from the effect shown in the preceding figures. Buffalo and Ithaca were both nearly in the middle of the shadow path, and are separated by a distance which is but little greater than the width of the shadow spot, so that for a short time the transmitter and receiver were nearly joined by darkness. Another striking feature about this record is that, unlike the preceding figures, the mean field and amplitude rise from 8.00 to 9.10 A. M., slowly at first, and then with greater and greater steepness. First contact of the eclipse at Ithaca was at 8.00 A. M., and so, disregarding minor fluctuations, the signal intensity from WGR rose approximately at the rate at which the sun's light was reduced. And after the middle of the eclipse, the intensity fell off at about the rate the light came on.

Leominster and WGY are both outside and to the north of the shadow path, altho the eclipse was approximately 99 percent complete at both transmitter and receiver. The Leominster record of Figure 7 shows a general lowering of mean field for

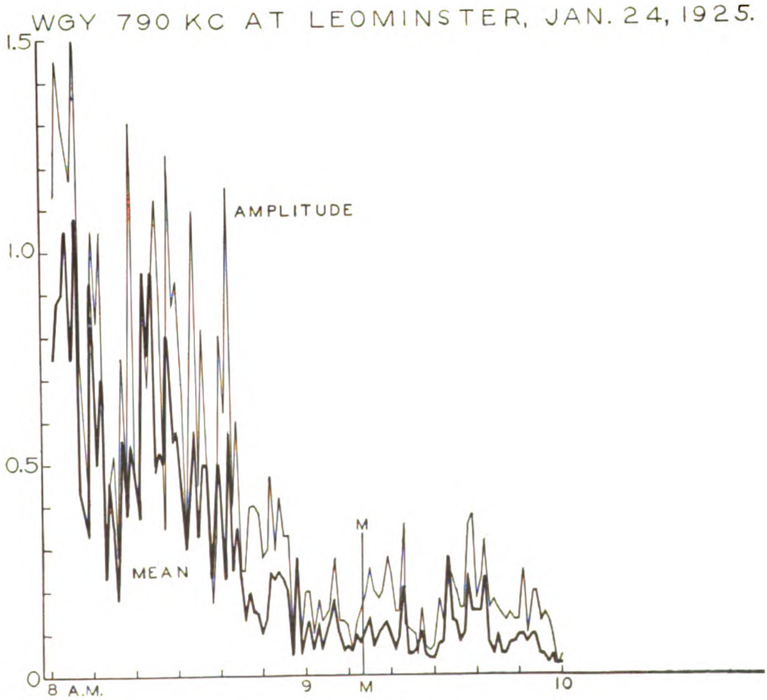


FIGURE 7

nearly half an hour at the middle of the eclipse, and superposed on this general lowering is a fall before and a rise after the middle which resembles that shown in Figures 4 and 5. But a still more marked fall and rise center on 9.30 A. M., so this record is not by itself conclusive.

Next to the Buffalo-Ithaca record of Figure 6, the reception of WGY at Hamilton, shown in Figure 8, is one of my most striking records. Altho the transmission path, as in the preceding figure, was wholly outside the shadow path, the fall and rise of both mean and amplitude curves are most strongly marked. The early portion of the Hamilton record shows a large night effect, which may be the explanation of the magnitude of the eclipse effect.

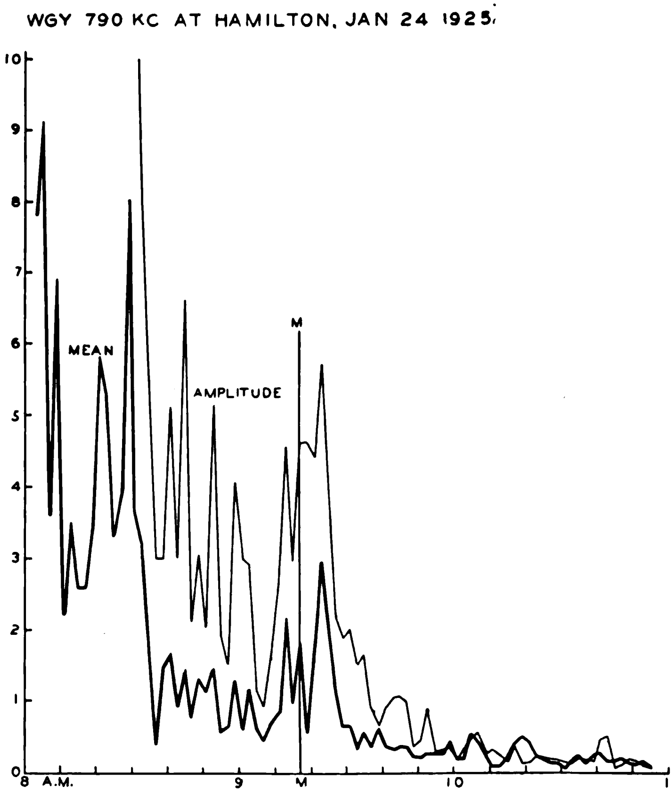


FIGURE 8

Figure 9, of WGY reception at Middletown, is not unlike the record shown in Figure 5. The eclipse effect is clear, altho a similar change of greater amplitude centers on 8.30 A. M. The

transmission line from WGY to Middletown, as may be seen by reference to Figure 1, is quite similar in its relation to the shadow path to the line joining WGY and Ithaca, so that if this is any criterion, the records should be alike.

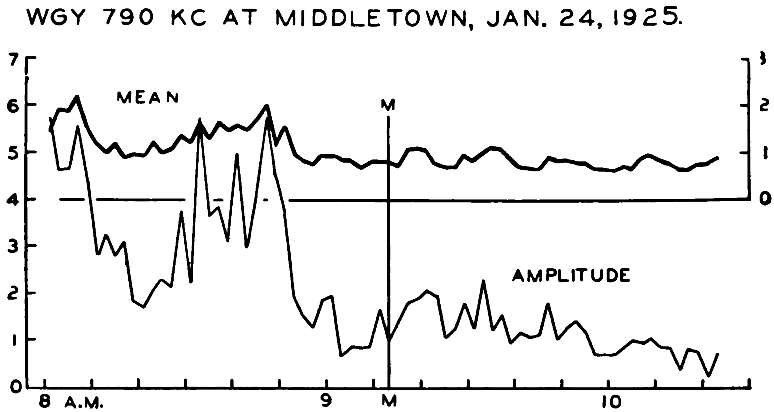


FIGURE 9

New York City reception from WGY, shown in Figure 10, exhibits a mean field which rises slowly from 7.30 to 9.22 A. M., and then slightly declines again. The amplitude curve shows a marked fall before totality, and a still more marked rise afterward, but similar effects appear at other points, as at 8.20 and 10.10 A. M. The transmission here is from a transmitter out-

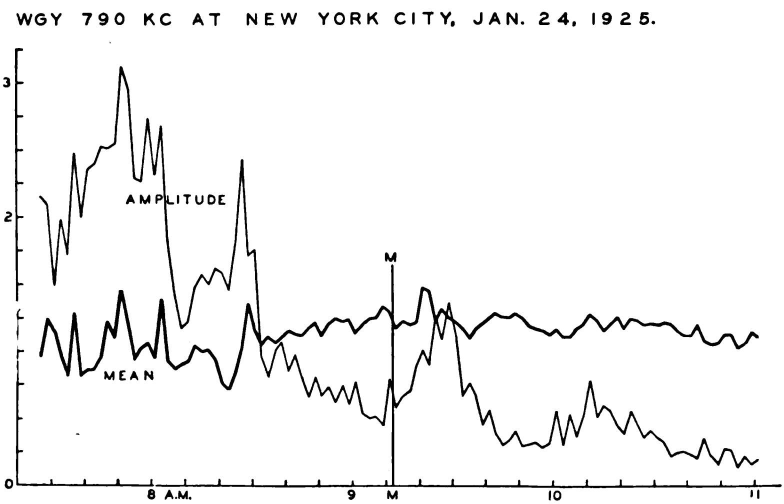


FIGURE 10

side and to the north of the shadow path, to a receiver inside but near the southern boundary of the path.

Rocky Point, the transmitter of the record shown in Figure 11, was nearly at the center of the path of totality, while the receiver at Washington was well outside and to the south. The record from beginning to end, save for the eclipse effect, is flat, and the fall and rise is unmistakable. It should be noted that this record is in ordinates which are proportional to field values, whereas in the other records which I have shown they are in field-square units. The 26 percent increase at the peak of this curve would, therefore, correspond, on the other records, with an increase of about 60 percent. Reception at such a low frequency is usually quite free from the short period fluctuations so common to the broadcasting band, and so altho this record was made from readings taken at five-minute intervals, a continuous record would probably have looked much the same.

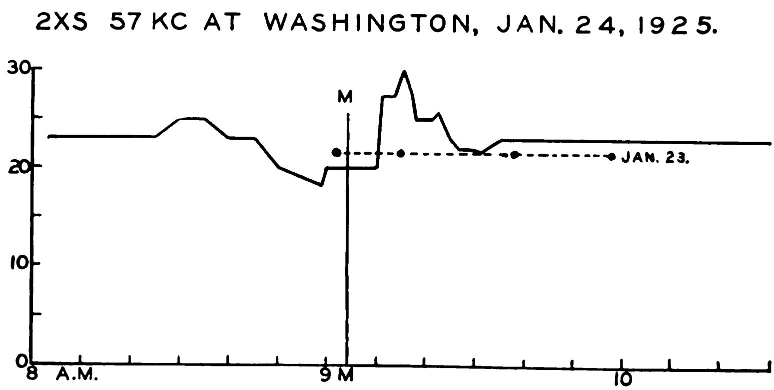


Figure 11

The record shown in Figure 12 is of WGY reception at Ithaca on the 26th, which at this receiving point most nearly corresponds with the eclipse morning. Altho slightly irregular, it would be difficult to find a pseudo-eclipse effect here.

WGY reception in New York City on the 22nd is shown in Figure 13. At this receiving point the morning of the 22nd most resembled the eclipse morning, altho it must be admitted that the resemblance is not very close. A pseudo-eclipse effect appears on this record at 8.35 and another at 8.55 A. M.

In Figure 14 I have shown Cambridge observations of WBZ on the 24th and 25th, which at this receiving point and for this particular station were similar mornings. The eclipse record

shows the fall and rise of both mean and amplitude curves, altho the peak values occur nearly at the middle of the eclipse, instead of later.

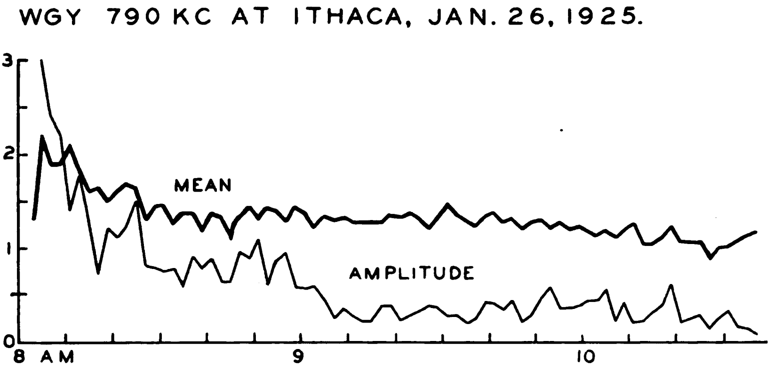


FIGURE 12

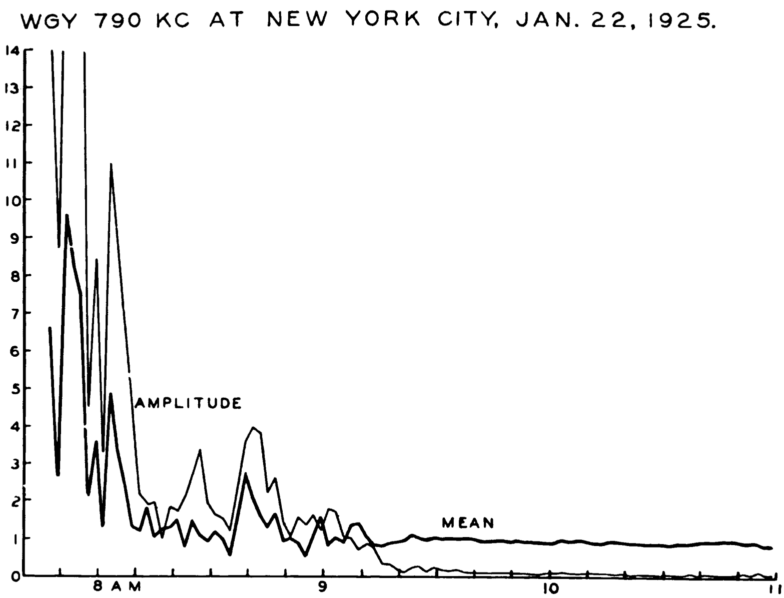


FIGURE 13

Figure 15, of Schenectady reception from WBZ, shows principally an increase centering on the middle of the eclipse. The smoothness of the curve before and after this increase is due to the fact that readings were taken at much greater intervals on these parts of the curve.

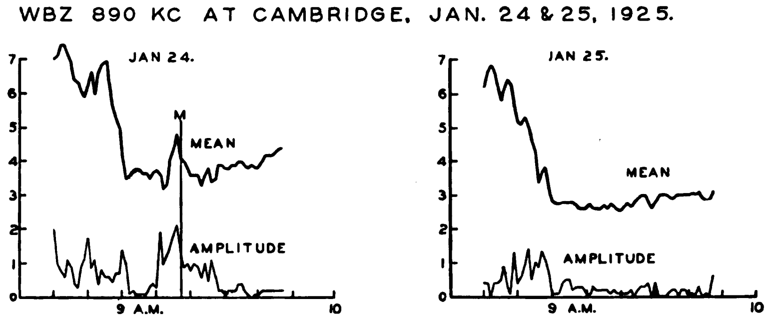


FIGURE 14

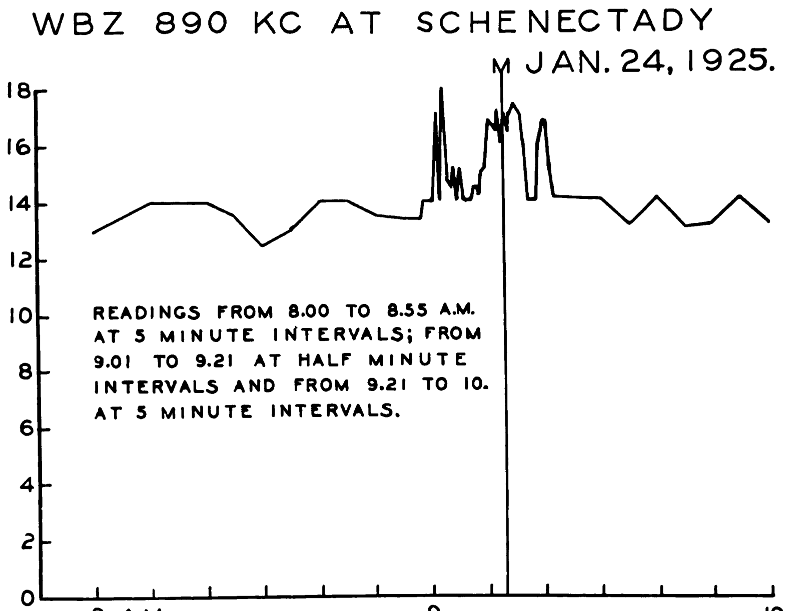


FIGURE 15

In Figure 16 I have attempted to bring out the general eclipse effect by taking four of my records, multiplying in each record the ordinates by such a factor as would bring the amplitude at the eclipse middle to approximately the same value, and than making an arithmetical average of the four. Altho this gives rather a shot-gun pattern for the points, I think that the dotted-line curves which I have drawn are fairly representative. This figure must not be taken too literally, and there is probably little significance in the apparent phase difference between the mean and amplitude curves.

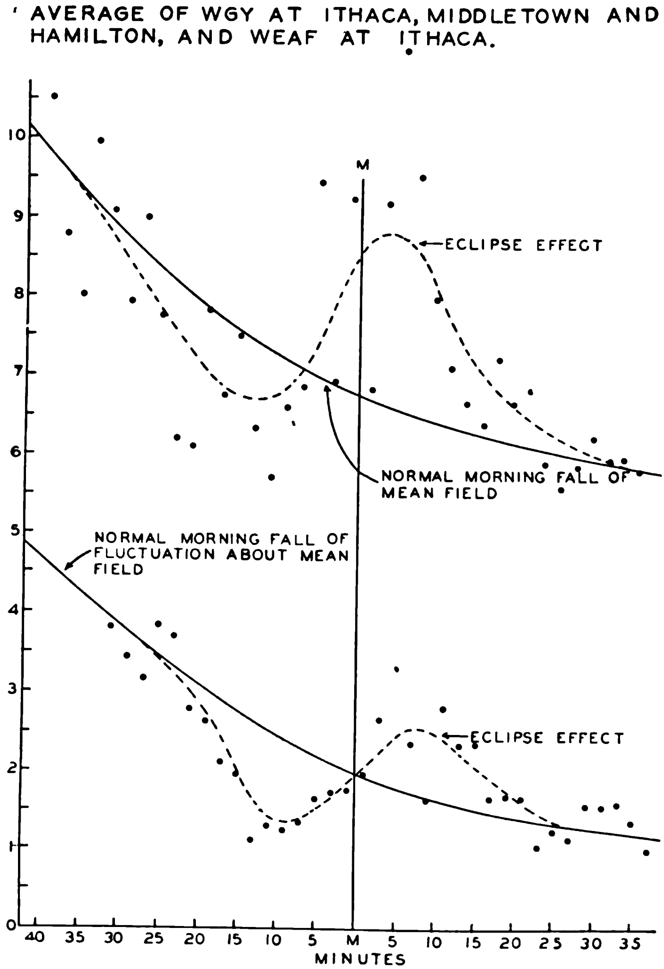


FIGURE 16

These morning records strongly remind one of an ebbing tide. There is the same incessant, periodic wash of waves, the same occasional high wave, and the same difficulty at any one moment in deciding whether the tide is going out or coming in. And it is this very irregularity in the recession which makes it hard to separate the eclipse effect from a mere coincidence. As I have pointed out above, there are few of my records which, taken by themselves, bear an unquestionable mark of the swift passage of the eclipse shadow. But when in record after record, both of different stations at the same receiving point, and of the same station at different receiving points, we find repeated the same

fall and rise, there can be little doubt but that we have found at least one of the effects of a solar eclipse on radio reception. And having found it, there can be little question but that its proper name is the Nagaoka effect. For in 1914, H. Nagaoka,[2] from purely theoretical considerations[3] announced that the intensity of signals should be weakened as an eclipse begins, pass thru a minimum, rise rapidly to a maximum, and then slowly decrease.

To explain satisfactorily the Nagaoka effect would require knowledge which we do not yet possess; detailed information of the pressure, composition, and ionization of our atmosphere at all levels, the exact effect of sunlight on this ionization, and finally just how the passing wave interacts with the ions. Probably the knowledge we have gained of the eclipse effect will find its immediate principal value in checking present-day transmission hypotheses.

The transmission of radio waves thru the atmosphere is, of course, a purely optical matter; we are merely working in the extreme infra-red as compared with the visible spectrum. From an optical standpoint the atmosphere is a highly complex medium, with equally complex effects upon any radiation which traverses it. As the atmosphere is of greatest physical, and therefore optical density at the surface of the earth, and rapidly becomes less dense as we go aloft, it is the equivalent of a prism with its base on the ground, and all radiation in a horizontal plane is therefore curved downward. This effect, which is practically limited to the first few kilometers of the air, accounts for a curvature approximately 25 percent of that of the earth, so that if this globe were four times bigger that it is, and had the same atmosphere, all horizontally emitted radiation would maintain the same level indefinitely, instead of going off at a tangent.

Because of the presence of gas atoms, and particularly on account of the fact that many of these atoms are ionized or electrically charged, the atmosphere must be regarded as an imperfect dielectric; optically considered it is something between a transparent insulator and a metal. This imperfect dielectric is subjected to at least two fields which have an optical effect; the earth's magnetic field, and the normal electrostatic field, which amounts to several hundred thousand volts across the whole atmosphere. Any radiation traversing the air will, there-

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[2] "Effect of Solar Eclipse On Wireless Transmission," H. Nagaoka, "Mathematico-Physical Soc.," Tokyo, Proc. 7, pages 428-430, December, 1914.

[3] "Effect on Radio-Telegraphy of Atmospheric Ionization," H. Nagaoka," "Mathematico-Physical Soc.," Tokyo, Proc. 7, pages 403-412, October, 1914.

fore, show both the Faraday and Kerr effects, that is, there will be a magnetic rotation of the plane of polarization of the waves along the direction of the magnetic field, and an analogous effect due to the electric field. That at least one of these effects—the Faraday—may be a large one for radio waves has very recently been suggested by Messrs. Nichols and Schelleng.[4]

It has been recognized for a long time that the optical properties of metals are quite different from those of transparent substances. A prism of glass bends light toward its base, while a metal prism often deviates light in the opposite direction, that is, it would appear that the light was propagated in the metal at a higher velocity than in vacuo. A very full discussion of this effect in metals has been given by Prof. R. W. Wood,[5] which leads us directly to our most recent attempt to explain the bending of waves around the earth.

One of the most promising of the many generalized explanations of the phenomena of radio transmission is the Eccles-Larmor[6] hypothesis. It appears to be the only one which has been numerically expressed without involving some grotesque amount or arrangement of ions in our upper atmosphere. Larmor first points out that true electrical conductivity in the air would act principally rapidly to attenuate or damp out the waves, and only secondarily to change their speed or course. But if the wave passes into a highly rarefied and only slightly ionized portion of the atmosphere, where the ions may be freely swayed to and fro by the passing wave-fields, there may be a change of speed without any damping of the wave. This change of speed or bending of the wave to conform to the earth's curvature will take place most effectively at some high level where the mean free path of the ions is so long that there may be many alternations of wave-field between two successive collisions with atoms or other ions. The ions at this high level will, therefore, sway freely under the alternating field, and interact without material dissipation of energy, increasing the velocity of the wave without either attenuation or scattering. Finally, as the effect is proportional to the number of ions present and to the square of the wave-length, each transmission frequency has a different level for its path, and very high frequencies may travel at quite a low level without absorption.

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[4] "Propagation of Electromagnetic Waves Over the Earth," H. W Nichols and J. C. Schelleng, "Science," Volume LXI, Number 1576, pages 288-290, March 13, 1925.

[5] "Physical Optics" 2nd Edition, 1914, pages 456-475.

[6] "Philosophical Magazine," December, 1924.

The picture we may make of such transmission is one of radiation struggling its way up from the transmitter, with severe absorption in the lower levels, particularly by day, but finally reaching a level at which it bends around the earth without loss, forming there what may be called a wave-sheet. As this wave-sheet spreads out in all directions around the transmitter, energy is showered down from it to the surface of the earth, again subjecting itself to severe attenuation in passing the lower strata. At very short distances, the receiver will be principally affected by waves which are transmitted more or less directly and at a low level; at moderate distances there will be a summation of both directly transmitted radiation and that dropping down from above, while at long distances the only material field at the receiving point is that coming down from the wave-sheet.

All this accords well with the observed phenomena of fading. At short distances there is but one path of transmission, with no interference and hence no fading; at moderate distances the fields due to the two paths are nearly equal in magnitude and so produce the maximum interference effect or fading, while at long distances the high level path is practically the only one to the receiver, and fading becomes much less violent. If the ionization which creates the wave-sheet were constant over considerable periods of time, there would be no fading, merely a diurnal change in intensity. At moderate distances from the transmitter there would be formed on the surface of the earth a stationary interference pattern, and a fixed receiver would remain in either a weak or a strong field, depending upon the accident of its position. At great distances there would be no interference pattern, and all receivers over very considerable areas would receive practically equal and non-varying signals.

But as a matter of every-night observation, we find rather rapidly varying signal intensity at moderate distances, and at long distances a less rapid fading. It would appear from this that the wave-sheet was far from smooth and constant, but was instead traversed by varying indentations and bulges, due to non-uniform and varying ionization. At moderate distances this would act to vary the intensity and path-length of the high-level route to the receiver, producing a strong and complex interference with the direct-path radiation. At the longer distances, where the directly transmitted wave is no longer an appreciable factor, the corrugations in the wave-sheet would produce plural path transmission down to the receiver, and fading is still ob-

served, altho different in character and less severe than that at moderate distances.

I have already had the honor of presenting before this INSTITUTE a large number of reception records[7] which fully show the change in fading with distance. It would naturally be expected that when a certain limiting distance was exceeded, there would be little or no change in the character of the fading, so that, save in the matter of intensity, there would be very little difference here in the East between reception from Chicago or from California.

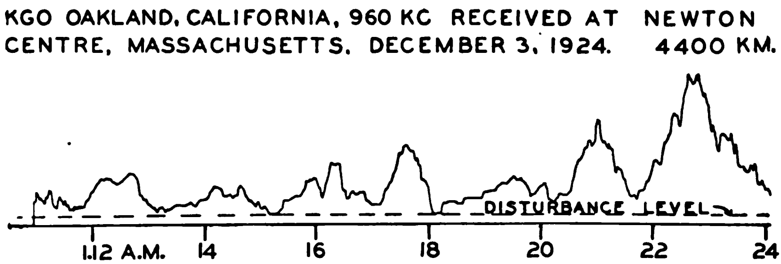

FIGURE 17

Figure 17, which is a reception record taken in Newton Centre, Massachusetts, from KGO at Oakland, California, shows a fading curve which is very much like that of Chicago stations of similar transmission frequency. This record is also interesting in that it shows a low disturbance level on that particular night; unfortunately this is not normal at my receiving point.

There is every reason to believe that the irregularities in the wave-sheet are in general small in area, and therefore individually affecting relatively small reception areas on the earth below. This is indicated by my repeated observations[8] of different fading curves for slightly separated receiving points. And if the wave-sheet is at or near the auroral level, the non-uniform appearance and rapid motion of the auroral streamers is perhaps an indication that the ionization at these high levels is also non-uniform and rapidly moving or varying. There is in fact some support for the conception of a moving series of irregularities in the wave-sheet, which would carry a more or less constant interference pattern over the ground below; the fading tests con-

[7] See these PROCEEDINGS.
[8] See these PROCEEDINGS.
[9] "A Study of Radio Signal Fading," "Scientific Papers of the Bureau of Standards," Number 476, 1923.

ducted by the Bureau of Standards[9] indicated something of this sort in the so-called "traveling curves" which they discovered.

It is always difficult to make a diagram of a generalized explanation, and altho I have attempted this in Figure 18, I trust no one will construe it too literally. For simplicity I have shown merely a cross-section of the radiation on one side of a transmitter. At short distances there is little or no fading, as there is but one path to the receiver. As the receiving point is moved further

NORMAL NIGHT TRANSMISSION ACCORDING TO ECCLES-LARMOR HYPOTHESIS.

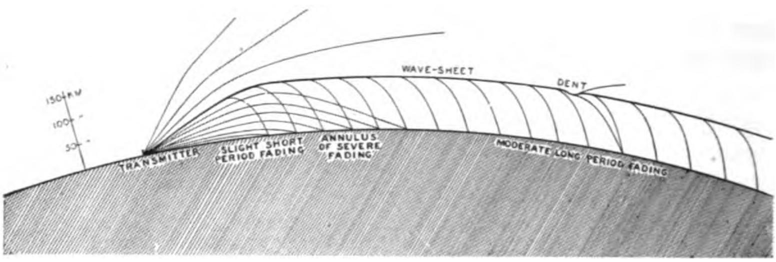

FIGURE 18

away from the transmitter, some energy begins to drop from above, and slight fading commences. I have quite uniformly found that at short distances—11 kilometers for example—the fading is not only of small amplitude as compared with the mean field[10] but shows a predominance of the shorter periods. This might be explained on the supposition, as indicated by the first descending line on my figure, that at short distances the indirect path to the receiver does not go all the way up to the level of the main wave-sheet, and is, therefore, subjected to different and perhaps more frequent irregularities that at the higher levels. As the distance further increases, the direct and indirect paths to the receiver deliver fields of approximately equal amplitude, and an annulus of violent fading is encountered, at a distance of between one and two hundred kilometers. Passing beyond this, the direct path transmission rapidly ceases to be a material factor, and less severe, longer period fading is found, due to varying or moving irregularities in the wave-sheet, which scatter energy up or down, depending upon the sign of the curvature, and so produce two or more paths down to the receiver, with consequent interference. One such dent is shown at the right of the figure; actually we must assume that the wave-sheet is full of them.

[10] My PROCEEDINGS article, Figure 11.

Coming back to the eclipse, it may at first sound fanciful to say that the moon's shadow dents in the wave-sheet, dispersing waves both up and down, and as this dent rapidly plows its way across the wave-sheet it accounts for the effects shown in my eclipse records. But, after all, this is no more fantastic to my mind than are some of our earlier and still tenaciously held transmission guesses, such for example as the conducting and reflecting Heaviside Layer. The moon's shadow diminishes solar radiation thru the air, the production of ions lessens, the ions in the shadow begin to recombine, and the total ionization falls. The wave-sheet within the shadow is now at a lower level than before, and has been literally dented in by the shadow.

It is evident that the distribution of wave-field on the earth's surface, resulting from the upward and downward deflection of energy at the edges of the traveling dent, will depend upon the relation of transmitter, shadow spot, and receiver. While, as I have shown above, the eclipse effect was generally a small one, yet, as in the instance of reception at Ithaca from Buffalo, a rather striking increase was found, which could hardly have escaped a broadcast listener's observation. At Easthampton, Long Island, a phonograph record of reception from WGY showed a marked decrease in intensity, beginning eleven minutes before totality, and lasting until twenty minutes after. A preliminary analysis of the "Scientific American" reports has clearly shown that the way in which the signal changed depended largely upon the position of the receiver with respect to the shadow path and the transmitter, and this may be summed up as follows:

When the transmitter and receiver were outside and on the same side of the path of totality, there was a gradual increase in signal strength, beginning about twenty minutes before the eclipse middle, and falling off again some ten minutes later. If the decrease before the middle of the eclipse has been too slight or gradual for aural observation, this effect, so far as the broadcast listener is concerned, appears in my records at Figures 8, 14, and 15.

When the transmitter and receiver were outside and on opposite sides of the path of totality, there was a decrease in signal strength beginning a few minutes before totality and lasting until well after. Mr. Robinson's galvanometer readings at Annapolis of reception from WGY show this effect.

When both transmitter and receiver were within the path of totality there was a relatively sharp increase in signal strength practically coincident with totality, which fell off rather rapidly

as the sunlight returned. Ithaca reception of WGR, shown in Figure 6, checks this very nicely.

In correlating my records with those of the "Scientific American" and other outside agencies, it now seems that one effect, suitably distorted to fit each case, will explain nearly all the observations. It is only necessary to assume that as the relative positions of shadow, transmitter, and receiver change, the relative magnitude of the Nagaoka fall and rise change also.

In Figure 19 I am once more showing something which must not be interpreted too literally. I have here illustrated in a general way how the simple Nagaoka effect will explain what are to the broadcast listener apparently opposite changes. In 1 the transmitter and receiver are both in the shadow path, and altho the signal first falls, and then rises, the accentuation of the rise makes it only part of the effect which is likely to be noted by the ear. In 2 the receiver is in the path, and the transmitter is outside, and with this arrangement my records show the fall and rise to be similar in magnitude. In 3 both transmitter and receiver are out of the shadow path, but on opposite sides, and this apparently accentuates the drop to such an extent that to ear observation it is the only effect.

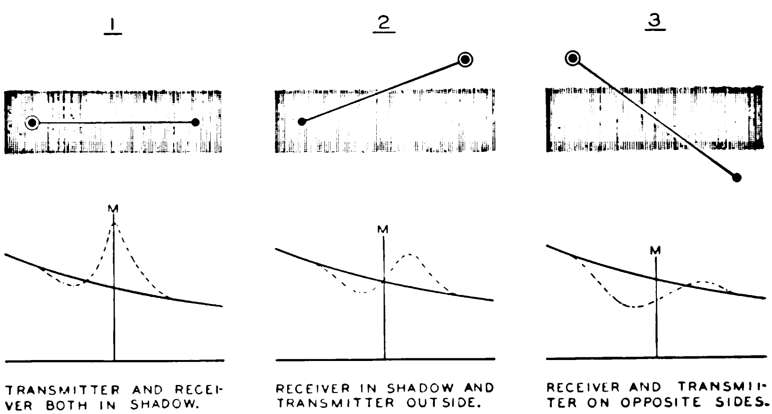

FIGURE 19

The many reports from observers of the eclipse broadcasting from station WTAT on the Atlantic Ocean were very kindly placed at my disposal by the Edison Light Company of Boston, and I have made a preliminary analysis of these reports, simply on the basis of whether or not the observer noted any change in intensity during the eclipse. Nearly half the reports stated that

there was no change in intensity, and nearly all of those noting a change found an increase reaching a maximum either at or shortly after totality.

I have embodied these results in Figure 20, the light circles indicating points where observers reported no change, and the dark spots points where an increase was noted. It is interesting to find that observers only fifty kilometers from the transmitter reported an increase during totality. As the figure shows, the greater part of the transmission was over water. Broadcast listeners as far away as Illinois, Canada, and Florida heard the broadcasting on the eclipse morning, but these distant listeners unfortunately failed to report whether or no the intensity varied.

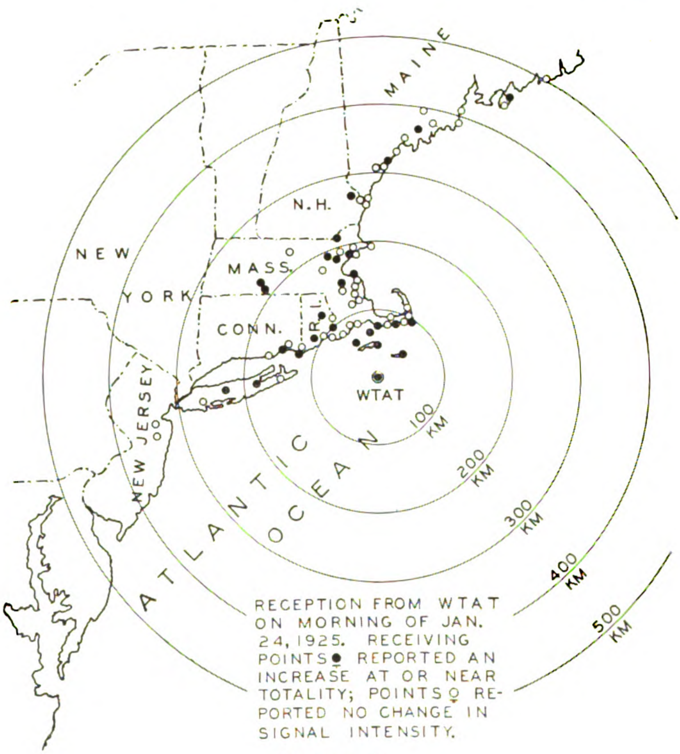

FIGURE 20

Our experiences with low-power broadcasting stations, which are normally on land and working overland, are such as to make this daylight transmission from a hundred-watt radiophone transmitter rather surprising. This particular transmitter had often been used in the past at points on shore in the vicinity of

Boston, but so far as my information goes it was then never heard by daylight at any such distances. Altho this is probably in no sense an eclipse effect, it may indicate the advisability of anchoring all of our coastal broadcasters off shore.

In conclusion I wish to express my appreciation of the cooperation so freely given by the Federal Telephone and Telegraph Company, the General Electric Company, the Westinghouse Company, the American Telephone and Telegraph Company, and the Radio Corporation of America. Without the transmission from their stations this work would have been impossible. I am deeply indebted to my assistants at Ithaca, Messrs. W. E. Bostwick, D. W. Exner, C. W. Garthlein, R. M. Holmes, C. J. Paddon and E. E. Zimmerman, and for the many courtesies extended to me by the Department of Physics at Cornell University, where my recording station was located. I am very grateful to Dr. Alfred N. Goldsmith, not only for his organization of the broadcasting stations for the eclipse schedules, but for the excellent recording work which was done under his direction. I also wish to thank my associate in this work, Mr. H. S. Shaw, who personally set up and directed the recording station at Middletown, and the many organizations and individuals which sent me valuable eclipse data and records. Finally, great credit must be given the Bureau of Standards, which not only took a very active part in the work, but also acted as an efficient clearing house for the distribution and collection of instructions and data.

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G. W. PICKARD (by letter): The records taken at the laboratory of the Radio Corporation of America in New York City, of the 4-megacycle transmission from 2XI at Schenectady, could not be reduced and replotted in time for insertion in my paper, so I therefore take this opportunity to present them.

The transmission from 2XI was interrupted every five minutes by keying, and in reducing these records I have taken the mean value for each five-minute period. In Figure 21, which is of reception on the morning of January 22nd, a general decrease is

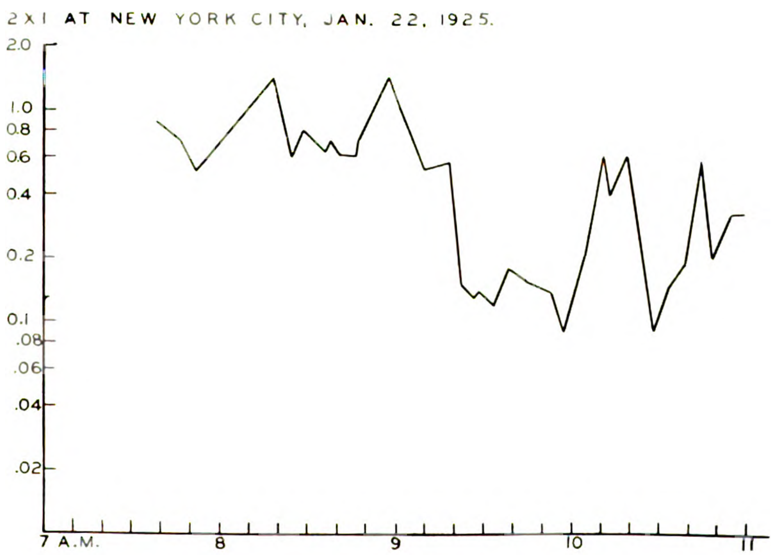

FIGURE 21

shown, the detector current changing over ten-fold, corresponding to a field decrease of over three times.

The morning of January 23rd, shown in Figure 22, is strikingly different from the preceding morning, the decrease in detector current being from 1,000 to less than 0.02; a range of over fifty thousand times, or a change in field of over two-hundred-fold.

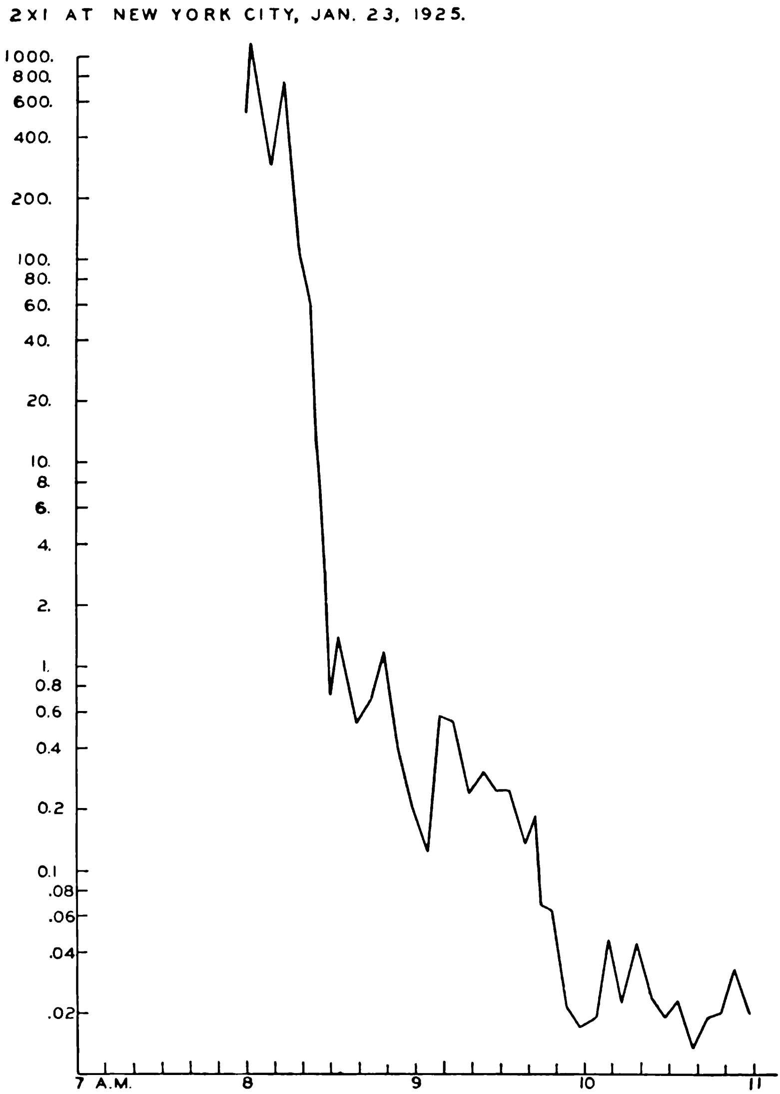

FIGURE 22

Figure 23 is the record of the eclipse morning, and at first sight it appears to be a complete reversal of the Nagaoka effect. But the rise to the peak value at 8.05 A.M. commenced sometime before the first contact, and so could not be an eclipse effect. From 8.58 to 9.46 A.M. the note of 2XI was inaudible, and over this period only the background was recorded. It would seem than this marked decrease was an eclipse effect, for it is not duplicated on any of the other four morning records.

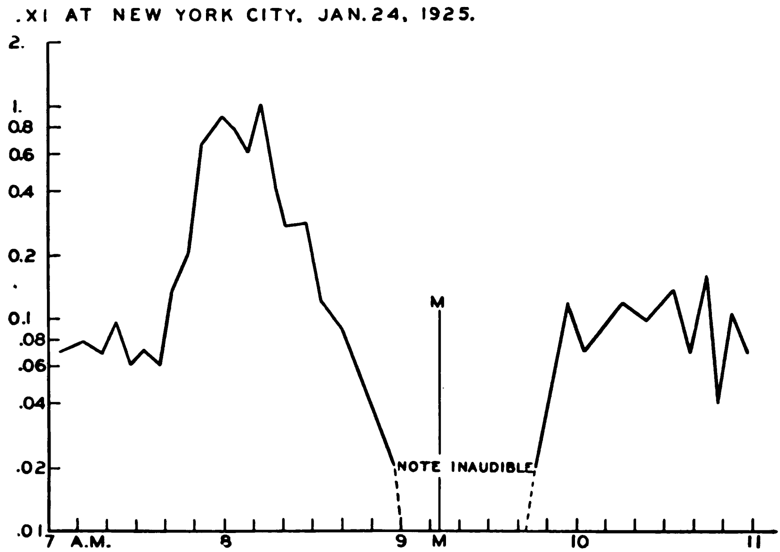

FIGURE 23

I have already mentioned the fact that WGY reception in New York City at 790 kilocycles gave flat records on January 23rd, 25th, and 26th. But, as appears from Figure 22, the morning of January 23rd was far from flat for 4-megacycle reception. So far as this single observation goes, one might say that it indicated transmission at a different level, agreeably according to the Eccle-Larmor hypothesis. Its real significance is that we need more such data before we can reach a definite conclusion.

Early in my paper, I said that one would naturally expect the eclipse effect to be similar to a night effect. I have recently obtained a series of sunset records at Newton Centre, Massachusetts, from WGY at 790 kilocycles, and as it seems relevent here, I give in Figure 26 an average of three of these records.

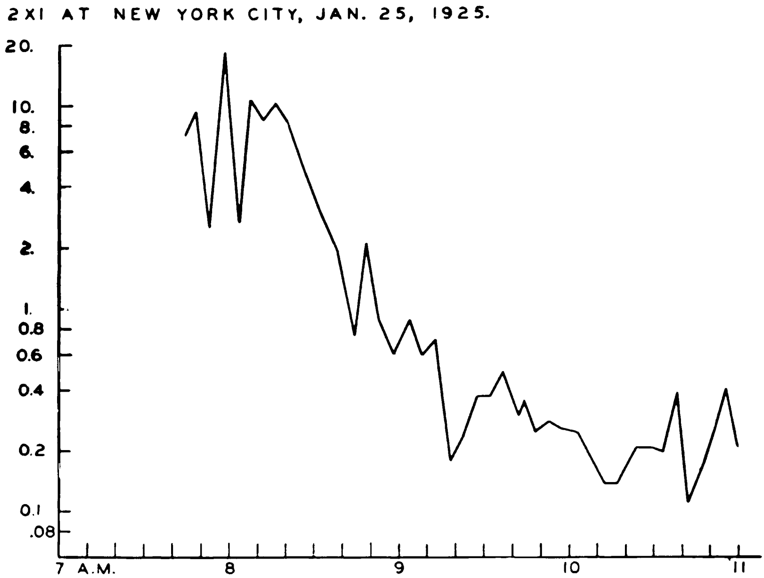

FIGURE 24

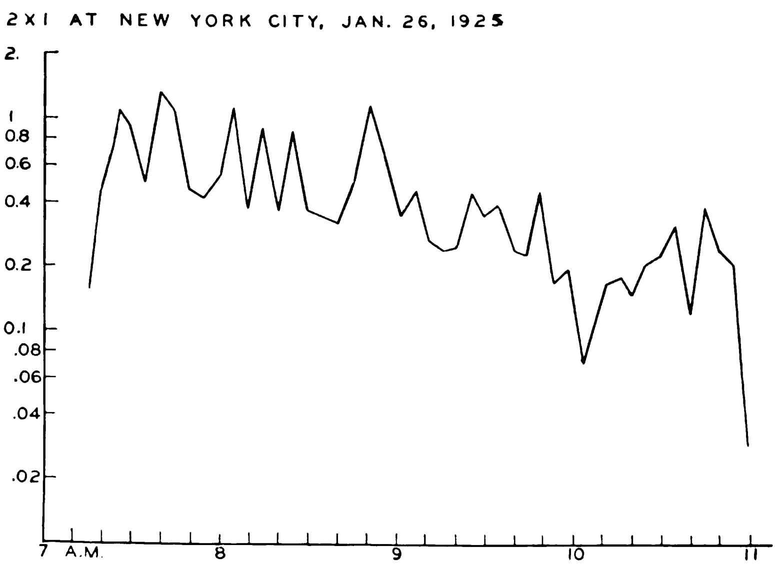

FIGURE 25

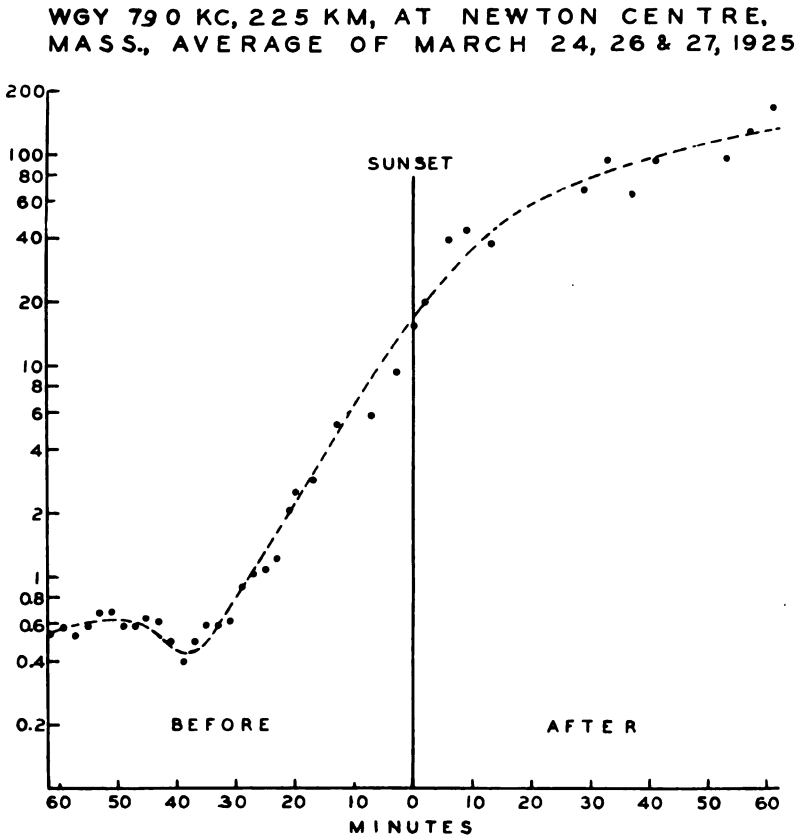

FIGURE 26

The dip at some thirty-eight minutes before sunset at Newton is not accidental; it appears on all the records, and altho not exactly at the same time on each, it was not wiped out by averaging. The rise from this dip to sunset is nearly a straight line, and as the ordinates of this figure are logarithmic, this means that the rate of detector current change was constant over this period.

TRANS-OCEANIC RADIO STATION—WARSAW, POLAND*

By

William G. Lush, Fred E. Johnston, and J. Leslie Finch

(Radio Corporation of America, New York)

History and General Features

By William G. Lush, Engineer in charge of Construction

In 1921, the reborn Polish State felt the necessity of having a system of international communication quite independent of the facilities of the surrounding nations; and it was decided by the authorities to erect a high power radio station for this purpose.

After a proper weighing of the kinds and capacities of the various forms of modern equipment it was decided to entrust the furnishing and erecting of the necessary apparatus, together with the engineering supervision of the enterprise, to the Radio Corporation of America.

A very considerable amount of work was required on the part of the Polish authorities and the executives and engineers of the Radio Corporation at New York in order to bring this matter not only into technical agreement with the very difficult conditions then obtaining for construction projects in Poland, but also that the project might be financed in such a manner as to be acceptable to both parties to the contract; which was finally signed on behalf of Poland on the first of August, 1921, by the Polish Minister at Washington, Prince Casimir Lubomirski and his commissioners and engineers, Stanislaw Arct, Hipolit Glivic, and Eugeniusz Stalinger; and on behalf of the Radio Corporation by Mr. Edward J. Nally, as president, and Mr. Lewis MacConnach, as secretary. The actual work in Poland was with the department making the contract, the Ministry of Posts and Telegraphs; headed during the greater part of the time by Mr. Jan Moszczynski, as Minister. The contract matters were handled by Mr. Stalinger as Chief Engineer for the Ministry of Posts, Mr. Stalinger being assisted in his work by Mr. S. Olszewski, the engineer appointed by the Ministry of Public

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Works to take care of the building and tower construction. The whole work was reviewed from time to time by a council of Ministers having supervisory powers. This Council consisted of the Minister of Posts and Telegraphs—chairman; the Minister of War, the Minister of Public Works, the Minister of Finance, and the Minister of Commerce and Industry.

In February of 1922 the Radio Corporation sent out Mr. Lush as Engineer in Charge of Construction and later Mr. Johnston joined the organization to take care of the receiving station, central office and the connecting lines with the rank of Assistant Engineer in Charge. He was followed by Mr. P. A. Baker of the General Electric Company and the latter's assistant, Mr. E. L. Marsh, who set up the transmitting set, power machinery, and antenna, except that the Diesel engine was erected by Mr. P. Derieckx of Carel Frères, of Ghent. Mr. Finch, who has charge of transmitter design for the Radio Corporation, came to Poland shortly before the completion of the station and took charge of adjustments, tuning and all final work prior to acceptance of the station by the Polish Government.

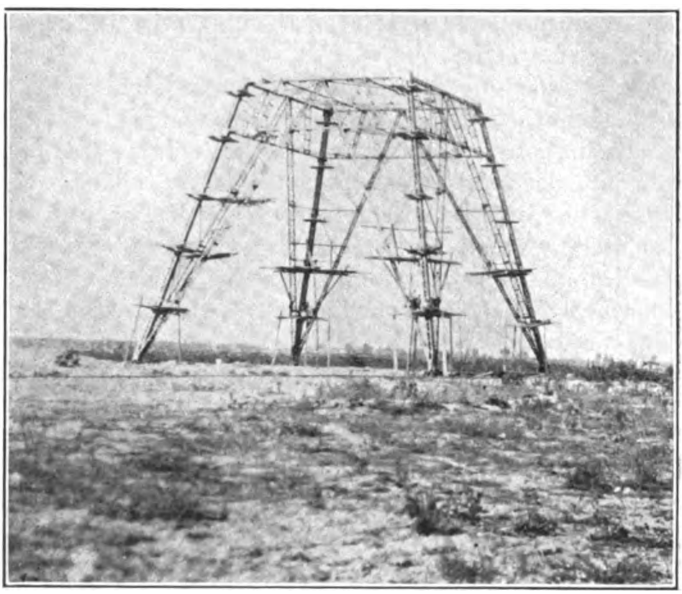

FIGURE 1—Base Section of One of the 400-foot Self-supporting Towers

The problem of site for the station had been satisfactorily solved by the selection of a very flat open stretch of land about

ten kilometers from the center of Warsaw. This land was free from obstructions on account of its previous use as a cleared area to permit gunfire from a circle of forts which surround the city but which are now dismantled. This site is very well suited for a radio station as it is almost entirely free from vegetation of any height and water is met at from one to three meters below the surface, thus giving excellent conditions for ground connections.

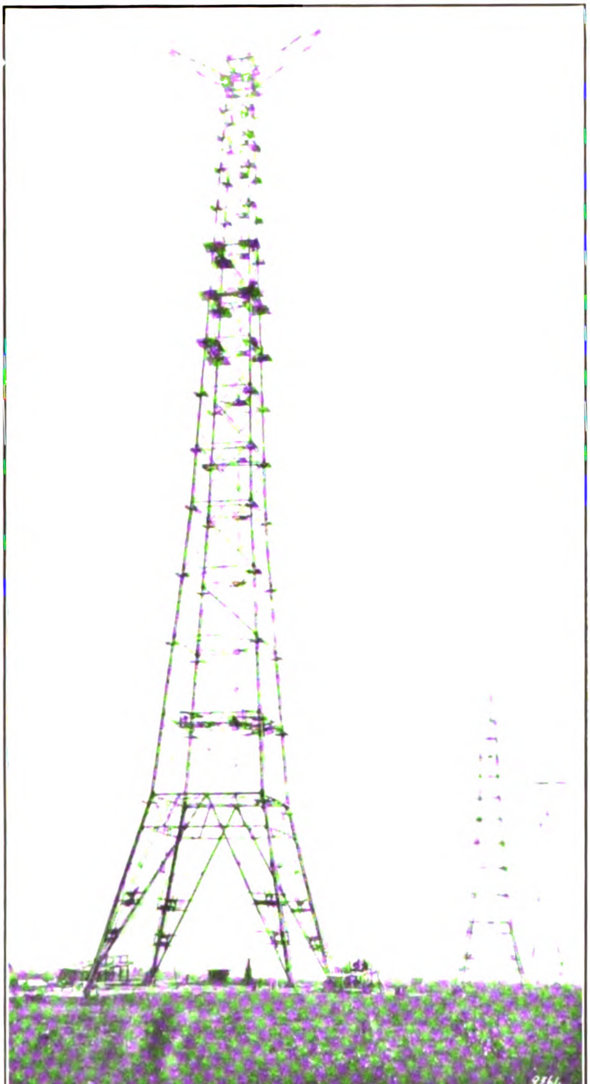

FIGURE 2—Three of the Towers in Construction Showing Cranes for Hoisting 75-foot Cross-arm Sections

The problem of shipments was a rather hard one. At the time of making the contract it was not certain as to just how much in the way of manufacturing facilities or the ordinary materials for electrical construction could be had from local or European sources and it was therefore necessary to make the shipments from the United States extremely complete.

The shipments came thru the port of Danzig and thence by the Polish State Railways to Warsaw, where the material was reloaded to a sixty centimeter military railway which reached the site of the station at Fort 2A.

The conditions in Poland were still somewhat disturbed, although steady improvement was noticeable. In consequence it was necessary for the Government to place guards over all shipments from the time that they left Danzig until they reached the ammunition casements at the fort. These casements were used as storehouses and proved most convenient for that purpose. In spite of the difficulties of transport and of training men to handle the expensive and unusual equipment, the loss by theft and breakage was very small.

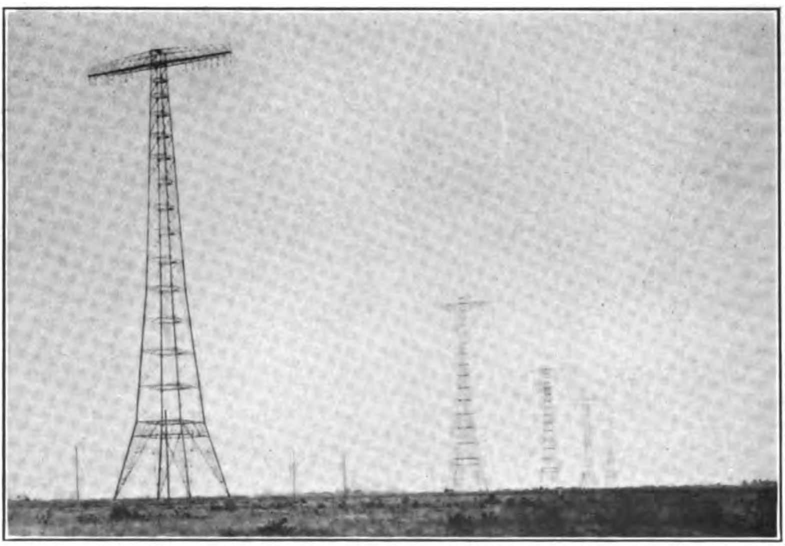

FIGURE 3—Line of Towers Showing Half of the Antenna

A problem of first importance was that of establishing an organization. This the contract provided that the Polish Government should do, but as the conditions regarding labor were much upset and as there were also present the difficulties which

are common to all governmental projects thruout the world, when the enterprise is of a novel sort, it was decided that the Radio Corporation would also undertake the handling of labor, reimbursement being made at actual cost. A special engineer, Mr. Waclaw Pogorzelski, was employed for this purpose and he was able to build up, as the needs of the work required, a labor organization entirely competent to perform the tasks in hand, in spite of the fact that the war had for the most part swept away the former supply of skilled workmen.

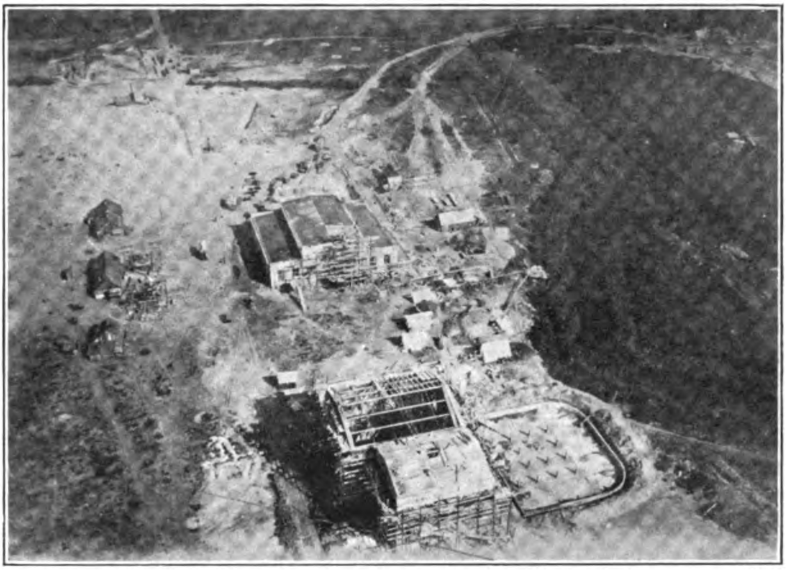

FIGURE 4—Aerial View of Radio Building and Power House Taken From Top of Tower Nearest Station; Taken During Construction of Buildings

Mr. Pogorzelski had spent some time in the United States and was, therefore, conversant with American as well as Polish methods, which contributed substantially towards the successful handling of this question. The Polish workman was found to be very skillful in carrying out a task once it was properly explained to him. While the rate of progress was not as great for any given amount of labor as with the highly trained personnel that we have in the United States, the results, under the circumstances, were very good and the men deserve much commendation for the way in which they endeavored at all times to carry out the recommendations and instructions of the engineers. It was found that the key to success in handling the labor was to keep

absolutely and to the letter all promises made to them in any way regarding conditions of working and promotion. With this established as a principle, we had a satisfactory amount of labor and were able to avoid almost all the strikes which occurred elsewhere during the period of construction.

The general working arrangement was to divide the construction into parts such as the laying of the ground wires, the installation of the alternators, the building of the tuning coils and the like. With any one division a gang of workmen would be assembled and these men would be shown their task by the engineer to whose activities this particular piece of work belonged, he giving his instructions thru the medium of an official furnished by the Government, who had a knowledge of the two languages. After a sufficient amount of work had been completed, one of the men would be selected to act as foreman and push the task to completion. The Polish officials were each able, of course, to supervise several of these activities, and while they were not all engineers, nevertheless each man was selected for his position with a view to his being a part of the permanent operating force later. This working principle was embodied in the contract and was found to be as successful in practice as any that could have been used.

FIGURE 5—General View of Completed Station, Looking West

The building of the station itself aroused great interest. The Pole is always interested in things which tend to increase his knowledge and improve his education, and it was noticeable that there were constant excursions to the work composed of all classes of society, from peasants and workmen, to the state officials.

The keenest interest was manifested by the students, the universities being hard at work to introduce modern engineering practices of all sorts into Poland.

While the materials for the electrical construction were supplied by the Radio Corporation, the Polish Government made its own contract for the buildings themselves, the boilers (Fitzmer and Gamper, Manufacturers, Sosnowiec, Poland), and the towers which were built by Rudski and Ska, of Warsaw. The local fabrication of the towers effected a most substantial economy.

The station buildings themselves are two in number, a power house and transmitting station proper. Both of these structures are built with thick brick walls covered with ornamental cement plaster on the outside and in the true architectural feeling of Poland, with heavy pilasters, entablatures, and heavy doors; the whole, however, expressing in shape and line the use for which the buildings are intended. The interior is beautifully finished in flat tints and devoid of any ornamentation except that furnished by well-proportioned spaces.

FIGURE 6—Part of Ground Laying Crew, Showing Trenching Shovels Used in Laying Ground Wire System

The transmitting station is a comparatively low building, having one story and monitor, with ventilators in the roof over the machinery hall. It is also furnished with a well-lighted office, store-room, baths, steam-heating plant and well-appointed machine shop, the latter equipped with an American lathe and the other necessary tools.

The power house is a tall building with rather straight lines and two pitched roofs, the one covering the boiler-room and the other the machinery hall.

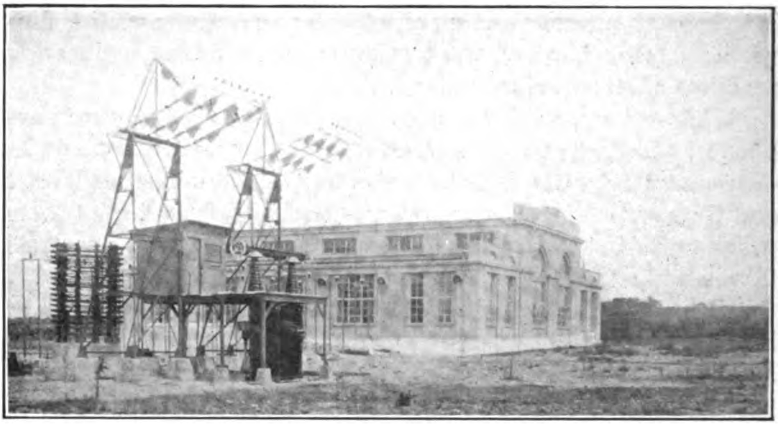

FIGURE 7—View of Radio Station and Terminal of One Antenna Wing, Station Tuning Coil, and Sleet Melting Transformers

The erection of the towers presented somewhat of a problem owing to the water in the soil, which is sand and clay, but this was successfully overcome with the use of concrete piles where the condition required them. Much of the concrete was mixed by hand, as contractor's machinery in Poland is not elaborate and everyone was used to the old method of dry ramming until after some time the force was sufficiently well trained to adopt the wet method for intricate pieces of work, such as the alternator foundations.

The staff quarters are much larger than is usual in American stations because under the Polish system of living each employee must be furnished not only with living rooms, but with a place to cook his food; even for those who rate only one room, a cook stove must be provided. Community dining-rooms, as in the United States, do not work well in Poland. Therefore, two large apartment houses were erected at the fort, giving ample and most comfortable quarters.

FIGURE 8—General View of Completed Station Looking South-east

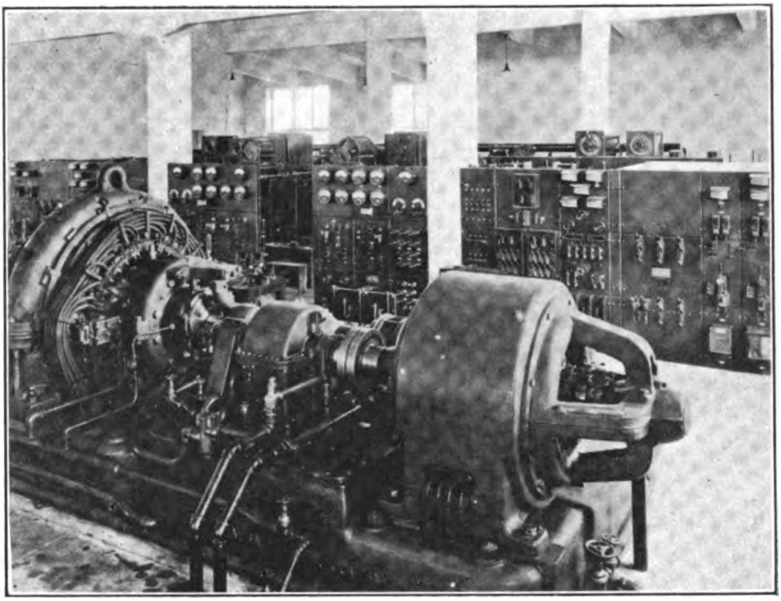

FIGURE 9—Interior of Radio Building Showing one Alexanderson Alternator and the Switchboard

It was necessary for the American personnel on this contract to be most particular in their dealings with everyone in Poland to avoid arousing, however unintentionally, any prejudice thru lack of acquaintance with the customs of the country, which has come to have an established formal etiquette as the natural outcome of the many years of history which lie behind it. Therefore, it was found at once that some of the rough and ready methods of action in common use on the projects in the United States would not apply and it was necessary to substitute, therefore, carefully thought-out ways of doing business, more in conformity with Polish ideas. One very happy result of this was that when something had been finally decided upon it was a fixed matter and could be depended upon to remain unchanged. It should be said at this point that the representatives of the Radio Corporation received at all times most courteous treatment at the hands of the Polish Government, that there is in Poland a sentiment very favorable to mutual relations between Poland and the United States and a very high degree of confidence that enterprises undertaken by reputable United States concerns will be carried out.

The station was opened for commercial traffic in October, 1923, and was formally turned over to and accepted, in Novem-

ber, 1923, by the Polish Government in the persons of the President of the Republic and the Minister of Posts and Telegraphs. The Radio Corporation personnel remained in Poland for some time after this formal opening to render such assistance as might be necessary in training and advising the Polish staff.

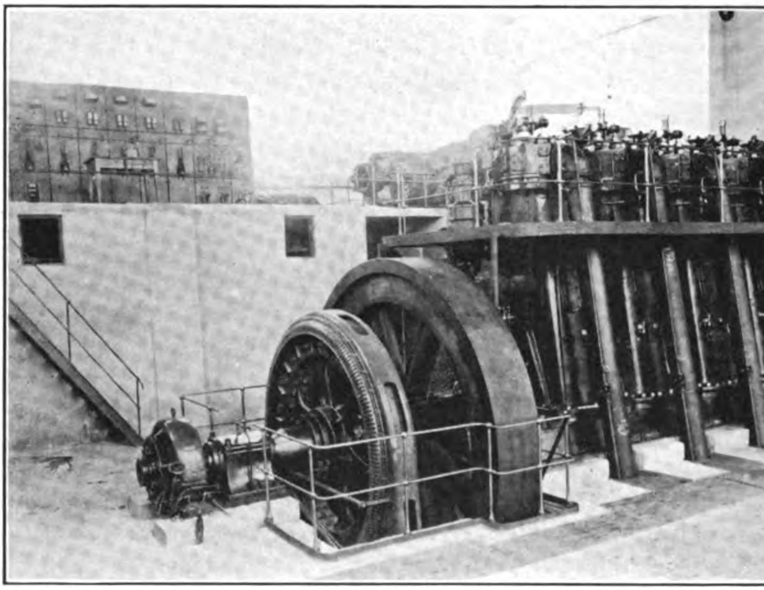

FIGURE 10—Interior of Power House Showing Diesel Engine and Steam Turbine and Switchboard in the Balcony in Background

FIGURE 11—Receiving Station Before Being Remodeled

In conclusion to these general remarks, the writer would express his belief that, provided proper consideration is given to local conditions, and to those with whom business is done, Poland does not present any insurmountable difficulties in the

carrying out of important contracts by any American enterprise which honestly intends to fulfil its moral and legal obligations.

Transmitting Station

By J. Leslie Finch, In charge of Transmitter Design

The transmitting end of the Warsaw station embodies the Alexanderson system, including the Alexanderson radio frequency alternator, the magnetic amplifier, and the multiple tuned antenna, and is essentially the same as is used in all of the high-power stations of the Radio Corporation of America. In the Warsaw station two complete 200-kilowatt alternator equipments are employed. These can be operated either singly, both at the same time or in parallel and will develop full power over a range of wave lengths from 18,000 meters to 21,000 meters. They are capable of operation in connection with the antenna at telegraphic speeds as high as 80 words (400 letters) per minute.

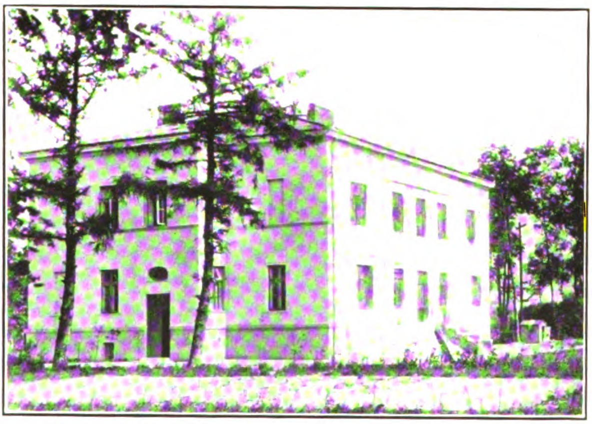

Figure 12—Receiving Station After Completion

The antenna is supported by ten towers of the self-supporting type. The towers are each 400 feet (123 meters) high and have cross arms at the top 150 feet (48 meters) long. These towers are placed in a straight line 1,250 feet (384 meters) apart, the radio building being located between the fifth and sixth towers.

There are 16 antenna cables running parallel with each other thruout the length of the antenna. They are supported by means of insulators and spreaders below the cross-arms of the towers. These cables are brought down to anchorages at each end of the

antenna and also in the center between towers 5 and 6. At this point they are broken and the two halves are entirely insulated from each other, except for removable jumper connections. This allows the antenna to be operated either as a whole or divided into two halves and each half operated separately. The total antenna has a capacity of about 0.08 microfarads and an effective height of 84 meters.

The tuning of the antenna is effected by means of inductors placed at the ends of each antenna half and adjacent to the towers at the intermediate points, ten inductors being used in all. Tne inductors are of the outdoor type, consisting of litzendraht cable supported in grooves in porcelain brackets which are in turn supported on a frame work of tubular porcelain uprights and copper rings. Each inductor has a total of 112 turns and an impedance of 18 millihenrys. This is sufficient to tune the antenna to a wave length of 23,000 meters. The ground connection to the inductors is made thru a remote controlled switch by means of which either of two predetermined taps may be selected. By throwing these switches one at a time and in various combinations, any one of a large number of wave lengths may be selected. Each alternator is connected in series with one of the inductors adjacent to the radio building. A variometer is also placed in this circuit by means of which small adjustments in wave lengths are made.

The ground system is of the conductive type and consists of wires laid in the earth and parallel to each other. They are laid at right angles to the line of the antenna and extending over its whole length. They extend to a distance on each side of the antenna of twice the height of the towers. Connection is made to these wires by means of buried busses and a system of overhead conductors which connect each of the antenna tuning inductors to the buried system at a number of different points.

The antenna insulators are of the porcelain tube type with galvanized steel end fittings. The units are each 53 inches (1.35 meters) long and are equipped with aluminum corona and rain shields at either end. Two of these units are used in series.

The total antenna resistance was found to be under 0.02 ohms. Thus it is possible to obtain a total antenna current of over 1,400 amperes when using the two alternators in parallel. This is the equivalent of 120,000 meter-amperes radiation. Using the antenna divided and one alternator with each half, a total of over 700 amperes is obtained in each half and two messages can be transmitted at one time. In practice the whole

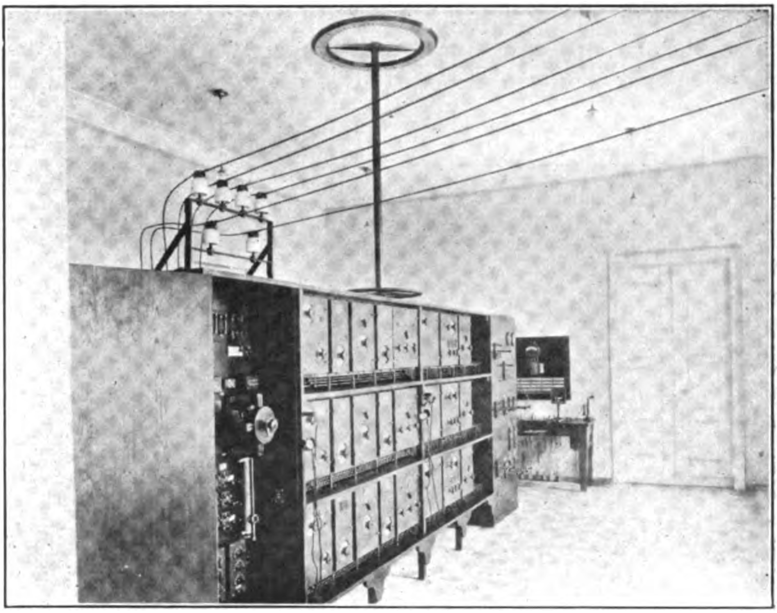

FIGURE 13—Interior of Completed Receiving Station Showing Three Long Wave Receivers Installed

antenna is usually used in connection with one alternator. In this way a total antenna current of 1,000 amperes is obtained from a power input to the antenna of 200 kilowatts.

At the wave length now employed, 18,350 meters, the antenna voltage with 400 kilowatts in the total antenna or with 200 kilowatts in either half is over 180,000. This voltage has been used without difficulty under all weather conditions. It will be noted that this voltage is much higher above ground than that employed in power transmission lines at the present time, since the 220,000-volt developments used only 127,000 volts, above ground. This voltage is maintained without visible corona. In order to maintain this condition it was necessary to protect all points otherwise subject to corona with suitable corona shields and to round off all sharp corners. The antenna cables used are 5/16 inch (8 mm.) diameter. One such cable supported singly at the same height as this antenna would develop corona at less than 100,000 volts. In the space between the towers the individual cables serve to shield each other and at the towers and near the earth the insulator fittings, spreader members, and the like, act as sufficient shielding except for the cables next to the towers. These were shielded by paralleling the antenna cables

for a short distance each way from the tower with a second similar cable. Corona on the downleads was prevented by constructing them in the form of a rattail or cage of 5 inch (13 mm.) diameter.

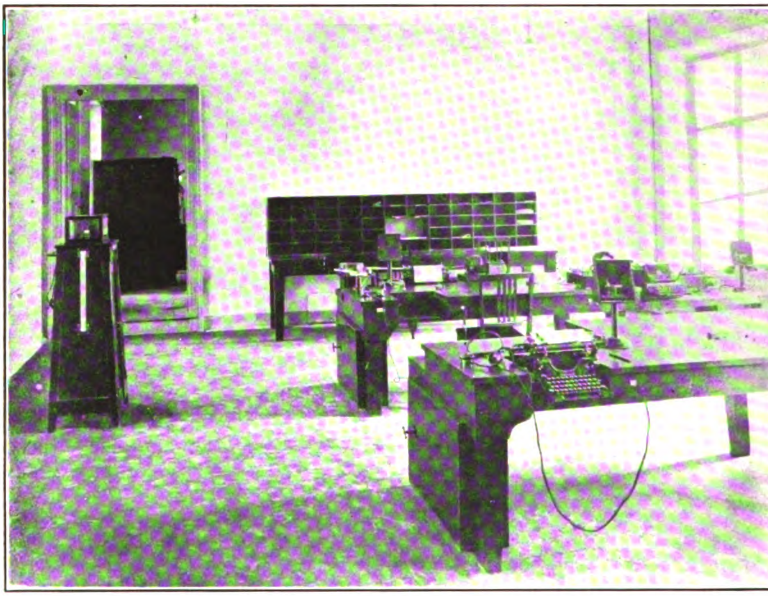

FIGURE 14—Interior of Operating Room of the Central Office

In order to remove ice from the antenna cables, which collects during sleet storms, they are arranged so they can be heated by passing current thru them from the same source of power as is used to drive the alternators. Provision is made at the radio building end of each antenna half to separate the antenna cables and connect the heating current leads. At the intermediate towers the antenna cables are kept separate for sleet-melting purposes by means of condensers. This is accomplished by suspending the antenna cables from the spreaders, which also serve as connecting conductors to the downleads, by means of short tubular insulators, inside of which are placed condensers of about one microfarad capacity, the end fittings of the insulators acting as terminals of the condenser. These condensers are essenially equivalent to direct connections for radio frequency currents and are equivalent to insulators for the power frequency sleet-melting currents.

The towers supporting the antenna are protected from over-

loads by suspending the antenna cables from the spreaders with weak links. In case of possible failure of the power supply for the station during sleet storms, the cables might become loaded with ice past their breaking point. Then when some of the cables had parted, the towers might be subjected to twisting stresses greater than they would withstand. This is impossible with the weak links, since they are adjusted so that they will give way before the cable tension has reached the breaking point.

The power supply at this station is furnished from a local plant adjacent to the radio building. Two generator units are provided, one a 500-kilowatt steam turbine and the other a direct connected 750 horse power Diesel engine generator set. Each unit generates power at 2,200 volts, 50 cycle, 2 phase. These units may be used separately, each to drive one radio frequency alternator, or they may be paralleled and used to drive both of the alternators. The operation of these two units in parallel has proven entirely satisfactory despite their widely different characteristics and the rapidly fluctuating load.

Receiving Station

By F. E. Johnston, Assist. Engineer in Charge of Construction.

The receiving station for the Warsaw transoceanic station is located in the outskirts of the village of Grodzisk, a small manufacturing and farming center, located about 19 miles (30 km.) southwest of Warsaw and on the main line railway between Warsaw and Cracow.

An old and much war-scarred school building was remodeled for use as the receiving station. It was large enough to permit apartments for the normal station staff being built in the same building. As is the case with most of the buildings in Poland, it has thick brick walls; these were covered with cement plaster on the exterior and the building presents an architectural appearance very much in keeping with the buildings at the transmitting station.

The wave antenna, as invented and developed by Messrs. Beverage, Rice, and Kellogg, and described in a paper presented before the American Institute of Electrical Engineers, is used. This antenna is 16,200 meters long and extends in a southeasterly direction from the station building, the center line of length being on a great circle bearing to New York and nearly at right angles to the direction of the transmitting station. In construction it resembles an ordinary telegraph line, consisting of two number 10 (2.5 mm.) hard drawn copper wires supported

by 20 feet (6 m.) wooden poles spaced 150 feet (46 m.). A 4-foot (1.2 m.) wooden cross-arm is attached to the pole tops at the ends of which the wires are supported on 22,000-volt glass insulators. The wires are transposed at every tenth pole.

It was found that most of the static arriving at Warsaw came from the south and southeast; thus due to its directional properties, the wave antenna gave a very great improvement in signal-to-static ratio over that obtained by the loop or loop and vertical method of reception. The station was also supplied with two outdoor type right angle loops, each 250 feet (77 m.) long and 9 feet (2.8 m.) above ground for the lower turns and 21 feet (6.5 m.) above ground for the upper turns. Each loop had twelve turns of number 12 hard drawn copper wire spaced 10 inches (23.6 cm.). An indoor loop was also provided, which was 8 feet (2.46 m.) square, wound with 100 turns of number 10 copper wire. These latter loops were installed for emergency use in case of failure of the wave antenna due to heavy sleet or storms.

Three sets of Radio Corporation standard long wave receiving sets were installed. These sets have a normal wave length range from 6,000 to 30,000 meters and are of unit construction, each unit being in a metal case which is well grounded. The parts of each unit are mounted on a panel of insulating material and a base so that all parts are within the metal case; the case is fitted with an aluminum hinged door thru which the controls work in such a manner that the door and dials of the controls form a complete shield. The back of the case, as well as the door, is removable so that all parts of each unit are readily accessible.

The filament and plate supply to all units requiring them are carefully filtered, both within each unit and at the source, to prevent disturbances entering the units via the battery leads. This precaution is necessary as all of the sets are operated from a common plate and filament supply. For normal operation the filament and plate supply is furnished by a 10-volt and 125-volt direct current generator, respectively, with storage batteries floating across the leads. The batteries absorb a large percentage of the commutator ripple and remain at full charge ready to take over the load in case of failure of the primary source of power.

The primary source of power is furnished by two gasolene engine-driven direct-current generators. These generators deliver 220 volts direct-current, and are located in a small building about 200 feet from the receiving building. For furnishing power to the sets, two motor generator sets are provided, each set con-

sisting of a 220-volt motor direct connected to a 10-volt direct-current generator and a 250-volt, 3-wire direct-current generator. These machines are located in a room adjacent to the receiving apparatus room.

The plate battery consists of 180 cells of lead type storage batteries arranged in three groups of 60 cells each: these cells are rated at 15 ampere-hour capacity. In normal operation two of these groups are used for supplying 220 volts to the plate of the oscillator in the synchronous detector unit and a tap taken at the connection point of the two groups serves to supply the 120 volts required by the amplifier units.

The filament battery consists of 8 cells of lead type storage battery arranged in two groups of 4 cells each, thus providing two 8-volt filament batteries. These cells are rated at 720 ampere-hour capacity and in case of emergency one group is sufficient to supply three sets for eighteen hours.

As no operating is done at the receiving station the output from the sets is put directly on to telephone lines to the central traffic office in Warsaw. Sufficient jacks are provided in a control panel for monitoring the signals and transferring signals from one line to another. For this transfer of signal, four special telephone lines were installed by the Polish Government. They differ from the normal telephone lines in Poland only that they are of bronze wire instead of the usual galvanized iron which is used extensively for telephone work in Europe.

Central Office

The central office is located in the Central Telegraph Office of the Ministry of Posts and Telegraphs in Warsaw. Complete telegraphic control of the station is centered at this office.

The four telephone lines from the receiving station as well as the three control lines to the transmitting station terminate here. Equipment for manual or automatic sending and receiving is provided, together with two radio receiving sets for monitoring the transmitted signal.

A system of local communication between the central office, the receiving station and the transmitting station is maintained over the tone or control lines by telephone, there being sufficient spare lines for this purpose. The necessary communication is thus quickly and easily handled and the personnel at the transmitting and receiving station need not be trained telegraphists, except as their other duties require this knowledge.

January 3, 1924.

SUMMARY: A brief history and description of the Warsaw Trans-oceanic Radio Station as constructed and installed for the Government of Poland by the Radio Corporation of America is given.

A technical description of the details of the system used is not given, as the system is similar in all respects to that in use in the United States by the Radio Corporation. Papers on this system have been given before this INSTITUTE and before the American Institute of Electrical Engineers by Messrs. Alexanderson, Beverage, and others.

THE APPLICATION OF THE X-L FILAMENT TO POWER TUBES*

By

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One of the chief problems in connection with the design of vacuum tubes has always been to secure a suitable material for making the cathode. Several properties are essential. The life should be long. The power required to heat the filament to the proper operating temperature should be as low as possible. The filament should not change greatly during the normal life of the tube. Also, for a given power consumption and total electron emission the filament should be comparatively long.

The fulfilment of the first two requirements, which are too often at variance with each other, usually means a relatively low operating temperature, the exact choice of which, for any one material, must often be a compromise between life and efficiency.

At the present time the only three materials which have been used to any great extent in power tubes are pure tungsten, oxide-coated platinum or platinum alloy, and X-L or thoriated tungsten.

The X-L filament as used in receiving tubes such as the UV-199 and UV-201-A radiotrons is already well known. Here its advantages have been very noticeable, resulting in marked improvements in operating characteristics with very low filament power consumption.

The same advantages appear even more prominently in the power tubes which have been designed with X-L filaments, some of which have been in commercial production for more than two years. The X-L filament is used in radiotrons UV-210, UV-203-A, UV-204-A, and UV-851, which are illustrated in Figure 1, and in a corresponding series of kenotron rectifier tubes. Only the radiotrons will be described in this paper however.

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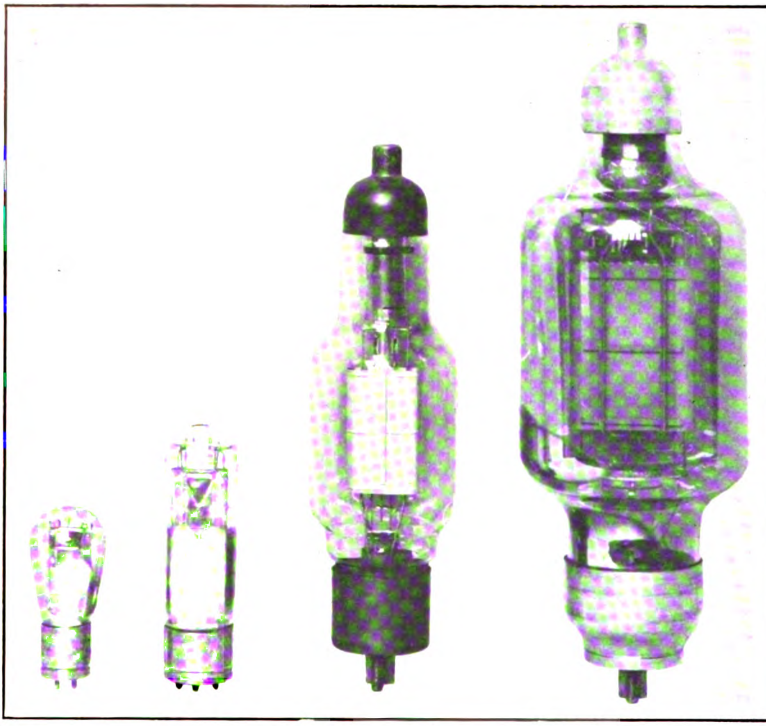

FIGURE 1— X-L Filament Transmitting Tubes

The normal operating temperature of the X-L filament is approximately 2,000° Abs. as compared with about 2,450° to 2,550° for pure tungsten. The emission efficiency of the X-L filament, however, is much higher than that of pure tungsten, as is well illustrated by the accompanying table, Figure 2, in which are shown the filament characteristics of several tubes of both types.

Tube	Type of Filament	Filament				Electron Emission	
		Volts	Amps.	Watts	Temp.	Milli-amps.	m.a. per Watt
UV-203	Tungsten	10	6.5	65	2,635	975	15
UV-204	Tungsten	11	14.75	162.5	2,480	750	4.6
UV-205	Tungsten	20	3.85	77	2,575	975	12.7
UV-210	X-L	7.5	1.25	9.4	2,000	700	74.5
UV-203-A	X-L	10	3.25	32.5	2,000	3,250	100
UV-211							
UV-204-A	X-L	11	3.85	42.5	2,000	4,900	115
UV-851	X-L	11	15.5	170	2,000	20,000	118

FIGURE 2

It will be noticed that the emission electron efficiency is not always the same for either type of filament. In the case of the pure tungsten filament the reason for this is that the choice of the operating temperature is a compromise between life and emission efficiency. For example, the UV-204 and UV-205 tubes are practically identical, except in their filament characteristics. The UV-205 (formerly known as the "Type P pliotron") was designed for war-time service where low filament power consumption was more essential than extremely long life. The UV-204, which was a later design, required 162 watts, but had a life expectation approximately 10 times as long as that of the UV-205.

The X-L filaments all operate at the same nominal temperature, but due to difference in lead losses, the emission efficiency is somewhat different in the various tubes.

In comparison with the UV-204 and UV-205, the UV-204-A gives an excellent example of the improvement effected by the use of the X-L filament. This tube retains the filament voltage of the UV-204 and the current of the UV-205, so that the power consumption is only 42.5 watts, which is only slightly more than one-fourth the filament power of the UV-204. On the other hand the life expectation is somewhat higher than that of the UV-204.

In the UV-204 the filament power is 39 percent of the total power expended in the tube. In the UV-204-A, the internal plate circuit loss being the same as in the UV-204, the filament power is only 14.5 percent of the total. This, of course, results in a lower plate temperature during operation.

There is another advantage gained by the use of the X-L filament, which is not at once apparent, but which is of very great importance. In designing a tube to be used either as an amplifier or as an oscillator, it is always desirable to make the space charge losses as low as possible, consistent with good mechanical clearances. This means that the effective electrode areas must be large and this can be accomplished only by the use of long filaments. Again a direct comparison between an X-L filament and a pure tungsten filament may be of interest.

The UV-203 tube requiring 10 volts and 6.5 amperes has a filament length of approximately 10 cm. The corresponding X-L filament tubes, the UV-211 and UV-203-A, requiring 10 volts and 3.25 amperes, have a filament length of approximately 17 cm. As a result of this increased filament length the mutual conductance of the X-L tubes is increased in about the same ratio.

In the UV-851 tube it was necessary to have a very large

cathode area in order to obtain the rated output of the tube at the relatively low plate voltage of 2,000 volts. For this reason the cathode was made up of four "V" filaments connected in parallel. For a given power consumption the total wire length of a filament is increased if two or more parallel branches are used instead of a single filament. This may be understood from a brief consideration of a few of the characteristics of tungsten filaments.

For constant temperature the filament current of a tungsten filament varies with the 3/2 power of the wire diameter.

$$\frac{I_1}{I_2} = \left(\frac{D_1}{D_2}\right)^{\frac{3}{2}}$$

For constant temperature the filament voltage varies inversely as the square root of the wire diameter.

$$\frac{V_1}{V_2} = \left(\frac{D_1}{D_2}\right)^{\frac{1}{2}}$$

For constant temperature and current the filament voltage varies directly as the length,

$$\frac{L_1}{L_2} = \frac{V_1}{V_2}$$

From these three relations it may be seen that the ratio of the lengths of two filaments of different diameter, but operating at the same temperature and at the same voltage, is equal to the cube root of the ratio of the two currents.

$$\frac{L_1}{L_2} = \sqrt[3]{\frac{I_1}{I_2}}$$

Applying this relation to the case of a cathode of constant power consumption, constant temperature and constant voltage, it will be shown that the total *wire length* increases as the cathode is broken up into more and more parallel filaments.

Let V = voltage across ends of cathode (that is, across all parallel branches)

n = number of parallel filaments

I = total current

$\frac{I}{n}$ = current in one filament

L = length of wire when only a single filament is used

L_n = length of single filament when n parallel filaments are used

Then $$\frac{L_n}{L} = \sqrt[3]{\frac{\frac{I}{n}}{I}} \text{ or } L_n = \frac{L}{\sqrt[3]{n}}$$

but since there are n filaments the total wire length becomes

$$L_{\text{total}} = \frac{n\,L}{\sqrt[3]{n}} = L\sqrt[3]{n^2}$$

For two parallel filaments the total wire length is increased 59 percent and for four the length is increased 152 percent.

There is a practical limitation to this method of increasing cathode area since decreasing the wire diameter too much will reduce the strength of the filament and increase the manufacturing difficulties.

The life of a tube having a pure tungsten filament is at best the burnout life of the filament. The operating temperature of this type of filament is so high that the tungsten evaporates at a fairly rapid rate, eventually resulting in the burnout of the filament. At the operating temperature of the X-L filament the evaporation of the tungsten is negligible and failure of the tube never results from a burnout except accidentally. The electron emission comes from a surface layer of thorium, and while this is constantly evaporating off it is also being renewed by diffusion of thorium from inside the filament. The theoretical life of the filament is ended then only when the supply of thorium inside the filament is completely exhausted. Also, since the electron emission depends only on the surface layer, it should be constant practically to the end of life.

On account of the reduction in diameter of a pure tungsten filament during its life, due to evaporation, it is best to operate at constant voltage and to use a voltmeter for regulation. The same does not apply to an X-L filament since there is no evaporation of tungsten, and either a voltmeter or ammeter may be used, altho a voltmeter is often more convenient particularly when tubes are run in parallel. Filaments should be operated at as near rated voltage as is possible, and, while an oscillator tube will give full output at considerably less than rated filament voltage, such operation is of doubtful advantage.

In order to maintain the active condition of the filament and the resulting uniform electron emission it is necessary to have an exceptionally good vacuum. This is accomplished by a thoro pump exhaust and the use of magnesium "getter" which deposits on the bulb giving the silvered coating characteristic of X-L filament tubes. Good electron emission during operation is an almost certain indication of a good vacuum, and any means taken to improve the vacuum tends to insure constant emission.

On account of the long life of the X-L filament itself, the life

of a tube containing this filament is largely dependent on the ability to design and manufacture a tube which will have the necessary vacuum conditions initially and will maintain them thruout life. In practice the failure of a tube often appears to be due to loss of electron emission; that is, the tube fails to operate satisfactorily because the emission has fallen off. Unless the filament has been subjected to excess voltage, loss of emission is seldom the true *cause* of failure, but rather the *result* of failure. That is, due to release of gas from the interior parts of the tube, often caused by overload, or due to leakage of air, the high vacuum is lost and the surface layer of thorium is destroyed, resulting in loss of emission.

In this respect the X-L filament is unique, in that the true filament life is so long that it is not a factor in determining the life of the tube.

In the series of radiotrons to be described later it will be noticed that many of the design features show the influence of this need for maintenance of a good vacuum over a long period of time. One of the commonest causes of air leaks is electrolysis of the glass in the seals which results in a crack in the glass or leakage along the leads. This is greatly accelerated by high temperature and in all of the X-L tubes the plate seal is placed at the coolest part of the bulb and is widely separated from the other leads. Also the distance between the filament and grid seal and the plate is made as large as possible. In the tubes operating at more than 1,000 volts, the so-called double end construction is used and the plate is supported entirely at the opposite end from the other electrodes. Also the grid lead is brought thru another separate seal.

Any power tube is apt to be subjected to short overloads in spite of reasonable care in operation and should be able to stand such conditions without permanent damage. An X-L filament will stand three times its normal voltage without immediate burnout and such treatment if not long continued does not seriously injure the filament as complete activity can be restored by a few minutes under normal conditions. Plate overload usually causes gas to be given off from the overheated plate or other parts of the tube and may cause a decrease in emission. In order to protect the tube against permanent damage from this source the X-L tubes are at present all made with molybdenum plates, which are heated to extremely high temperatures during exhaust and thereby are freed of most of the gas contained in them. The high temperature of the plate also heats the other

internal parts of the tube hotter than they should ever become during operation and reduces the chance of their giving off gas later on.

The areas of the plates are relatively large and the surfaces are treated so as to radiate heat easily, so that in operation the plates should never run hotter than a dull cherry red color.

In the case of a very severe plate overload which has liberated considerable gas, the electron emission may be reduced temporarily, but this can almost always be restored by 15 minutes of operation with plate voltage off. This action may be accelerated by raising the filament voltage to about 15 percent above normal. After the filament is reactivated, the tube should be put into operation again under conservative plate load and the gas which has been given off during the overload will usually clean up. This applies only when the gas has been given off internally. An air leak invariably results in complete failure of the tube. If a glow appears in the tube its color will usually show whether the gas has been given off internally or has leaked thru from the outside. A glow which is distinctly blue is usually an indication of gas given off from the electrodes while a purplish or pink glow indicates an air leak. If the leak is sufficient to raise the internal pressure approximately to atmospheric pressure there will be no glow but the filament will give off a white, powdery smoke and will soon burn out. This, of course, is characteristic of both X-L and pure tungsten filaments.

It might seem that the electron emission of an X-L filament, coming as it does from a constantly changing surface of thorium, would be more or less erratic and unstable. This, however, is not the case. In a properly operated tube the emission as well as the other characteristics are perfectly steady. Full emission occurs as soon as the filament comes up to temperature and there is no difficulty whatever in starting oscillations when the tube is used as an oscillator. Also there are no harmful secondary emission effects.

Several of the X-L filament tubes already mentioned will now be described in more detail.

Radiotron, Model UV-210, is the smallest of the group. The external appearance and the internal construction are illustrated in Figure 3. The electrode structure is more rugged than that of some of the earlier transmitting tubes and the electrodes are rigidly supported at each end. The bulb is of the so-called tipless type.

This tube has a nominal output rating of 7.5 watts. The

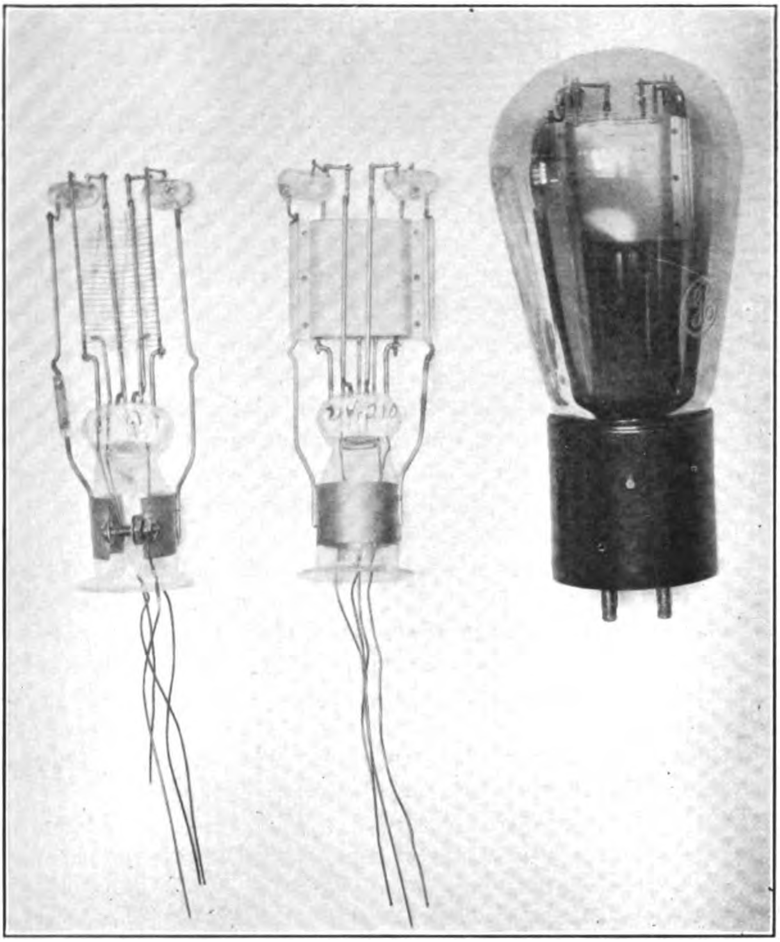

FIGURE 3—XL-7 1/2 Watt Transmitting Tube

normal plate voltage is 350 volts and at this voltage the rated output can be very easily obtained with a plate current of 60 milliamperes. If care is used in the construction and adjustment of the circuit, a much greater output may be obtained with the same input. The grid voltage-plate current and the plate voltage-plate current characteristics are shown in Figures 4 and 5.

The amplification constant of the UV-210 is approximately 7.5 and the internal plate resistance at 350 volts and zero grid is approximately 3,500 ohms. The corresponding mutual conductance value is 2,150 micromhos. It should be understood that these figures are useful for comparison purposes only, since under normal conditions the grid voltage is not zero.

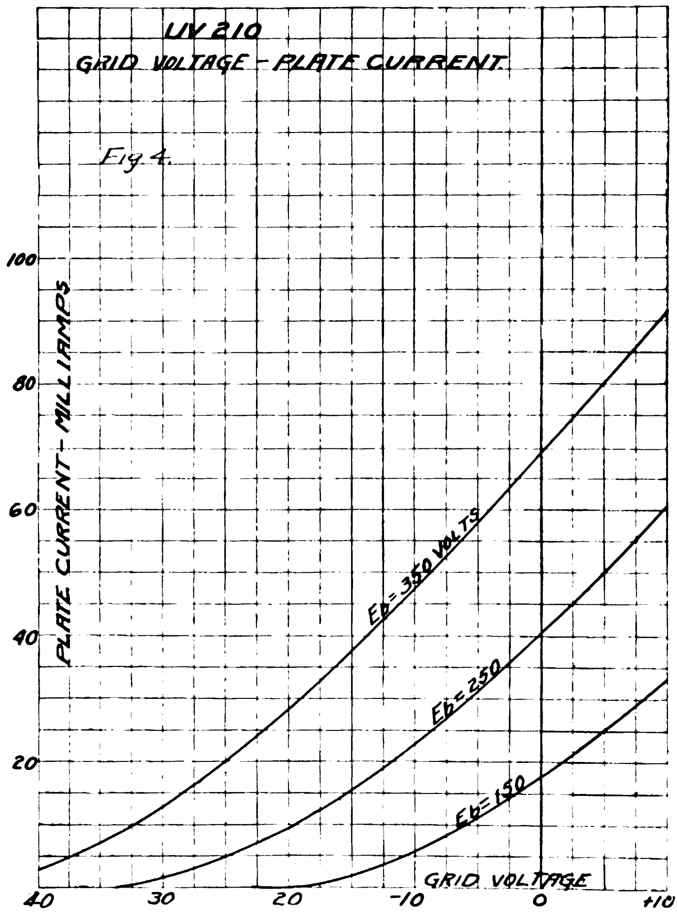

FIGURE 4

In addition to its use in transmitting sets as oscillator, modulator or power amplifier, this tube has been used to some extent in commercial receiving circuits where the service is continuous and where the power output is unusually high, requiring plate voltage and plate current conditions nearly as severe as in transmitting circuits.

The next larger X-L filament power tube is the 50-watt size, shown in Figure 6, which is made in two types, the UV-203-A and the UV-211. These two types differ only in plate impedance and amplification constant, and are exactly the same in external appearance, filament rating, and plate voltage. The UV-203-A is intended for voltage amplification or for amateur and experi-

mental use as an oscil ator, while the UV-211 is designed for oscillator, amplifier and modulator use in commercial sets.

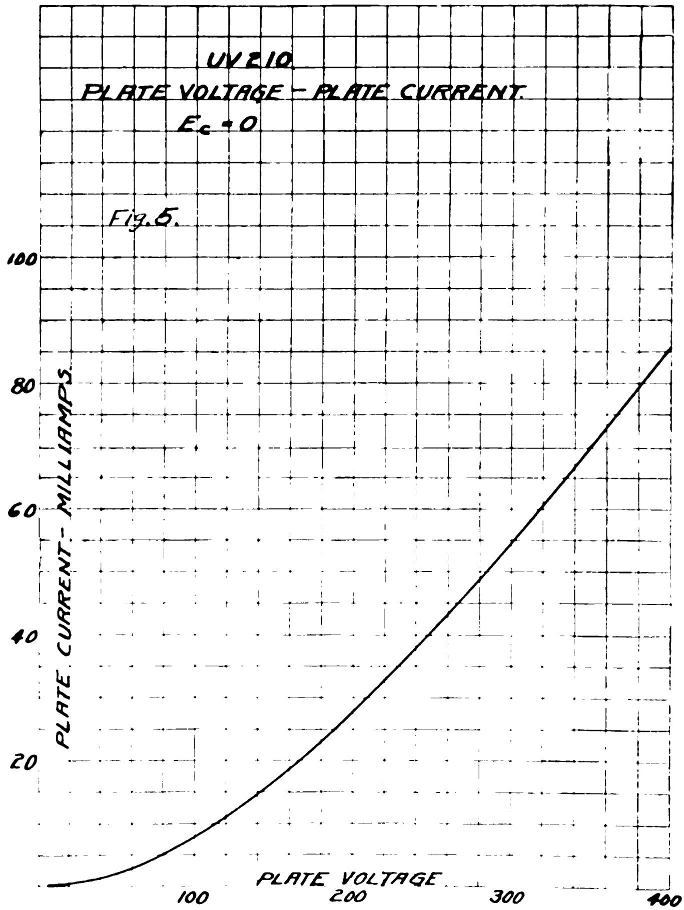

FIGURE 5

Both tubes have a filament rating of 10 volts and 3.25 amperes. The normal plate voltage is 1,000 and the maximum plate power dissipation is 100 watts when a tube is used as an oscillator. When it is used as a modulator or amplifier, the continuous dissipation is limited to 75 watts.

The UV-203-A supersedes the UV-203, and a few comparisons between the two tubes plainly illustrate again the advantages of the X-L filament and the general improvement in the later vacuum tubes of this size. The UV-203 required 10 volts

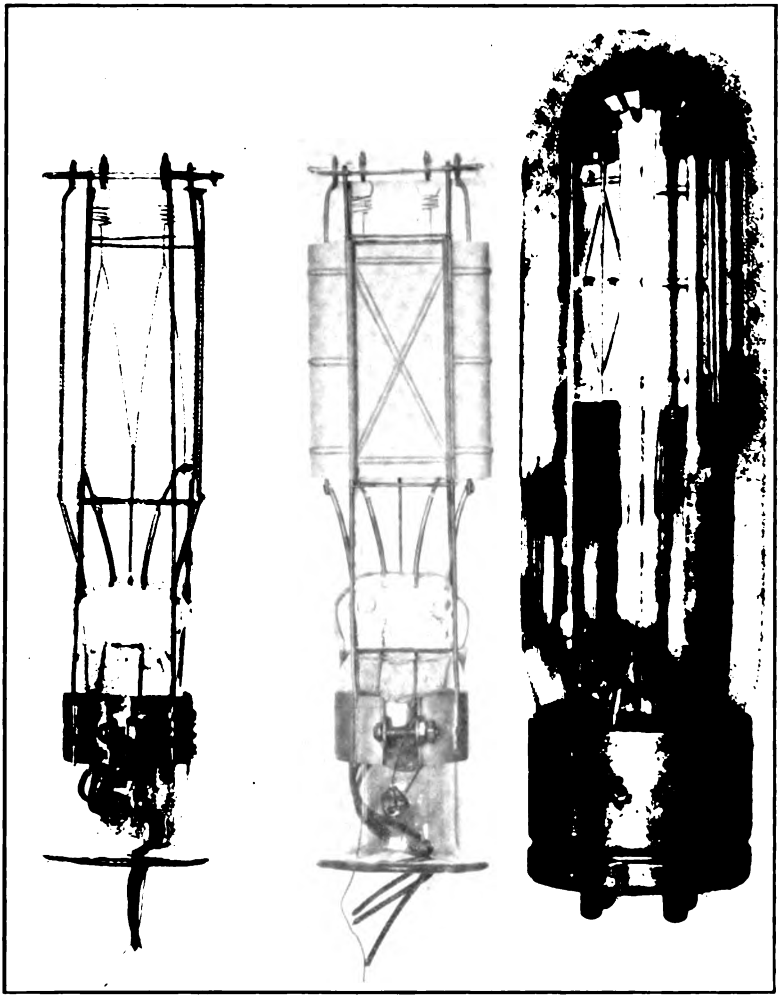

FIGURE 6—XL 50-Watt Transmitting Tube

and 6.5 amperes or 65 watts for filament heating, while the UV-203-A requires only one-half this amount. In spite of this, the electron emission of the UV-203-A is twice as great. The benefit derived from high electron emission and the improvement in characteristics resulting from the large electrode areas are illustrated in Figure 7, which shows the relation between filament voltage and antenna current for the UV-203-A and the UV-203 tubes in a typical transmitting circuit. The filament power is also shown for comparison. It will be noticed that the UV-203-A gives a higher output and that this output is obtained at one-

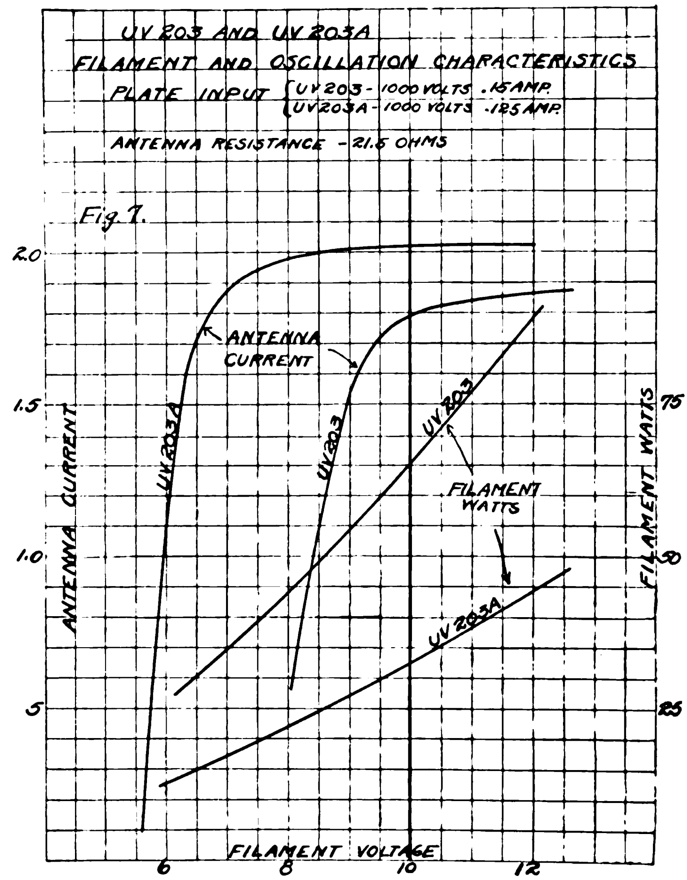

FIGURE 7

half the filament power and at one-sixth less plate current than is taken by the UV-203. At the normal operating point, the UV-203-A gives over 20 percent greater oscillation efficiency than the UV-203.

The UV-211 gives very similar results as an oscillator, but due to lower plate resistance it is much more satisfactory as a modulator or amplifier than is the UV-203-A. The lower impedance is very important where undistorted amplification is desired, as relatively large grid voltage swings may be employed without the grid becoming positive.

Figure 8 shows the plate voltage-plate current characteristics of the UV-203-A and UV-211, and Figure 9 the grid voltage-plate current curves. Also for comparison the UV-203 characteristic has been drawn on Figure 9. The steeper slopes of the

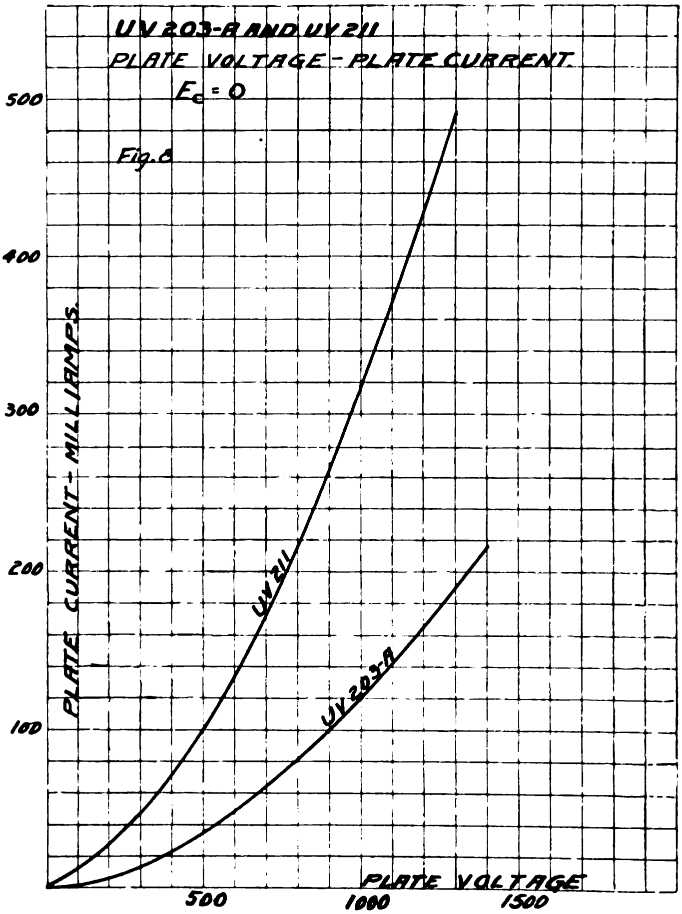

FIGURE 8

UV-203-A and UV-211 curves plainly indicate the superior characteristics of the X-L filament tubes.

The method of safeguarding the UV-203-A and UV-211 tubes from plate current overload may be of interest. The plate resistance of the UV-203-A is so high that if the tube should stop oscillating or lose its negative bias, the plate current would not greatly exceed the normal value when oscillating. Thus, the tube is self-protecting in this respect. On the other hand, the plate resistance of the UV-211 is so low that the plate current at zero grid would soon injure the tube by overheating of the plate. However, here the increase from the normal current during oscillation to the current at zero grid is sufficient to allow

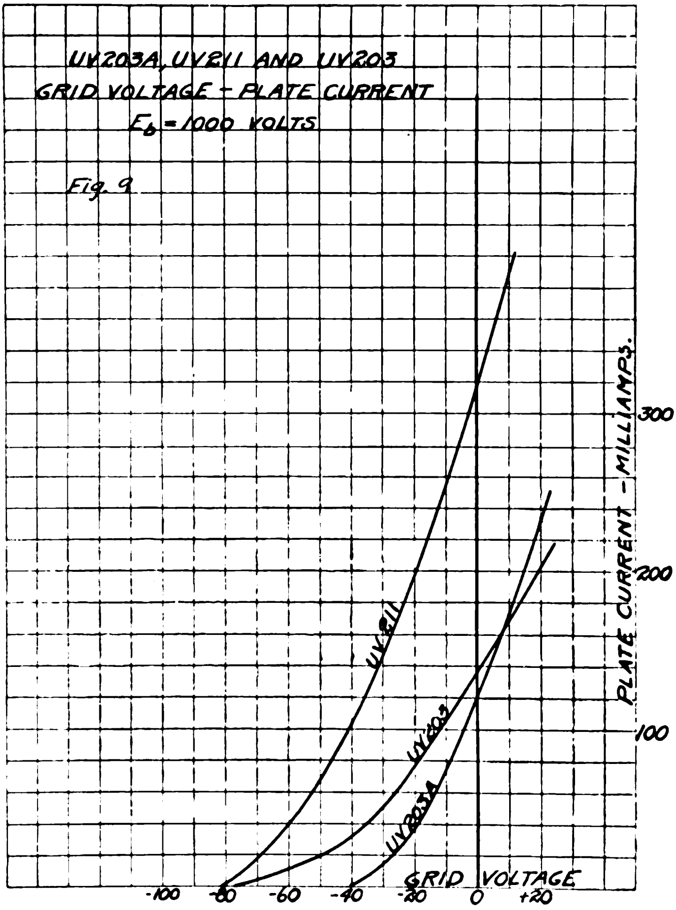

FIGURE 9

the use of a fuse or circuit breaker which will immediately open the circuit and thus protect the tube.

At the normal plate voltage of 1,000 and at zero grid voltage the UV-203-A has a plate resistance of 5,000 ohms, an amplification constant of 25, and a mutual conductance of 5,250 micromhos. The UV-211 has a plate resistance of 1,900 ohms, amplification constant of 12, and a mutual conductance of 6,300. A comparison of these mutual conductance figures brings out the fact that in two tubes of similar design, a slightly better mutual conductance usually is obtained in the tube having the lower amplification constant.

The mechanical construction of the X-L 50 watt tubes is much stronger than that of the UV-203. As in the UV-210 tube, the

electrodes are fastened together at the upper end. The plate is mounted on four support rods and has four wings which serve not only to make the electrode more rigid but also to dissipate heat. In this tube, helical springs are used to maintain the proper tension on the filament when lighted and also to protect it from shocks. This is necessary for tubes of this size or larger, while for smaller tubes the usual anchor wire has adequate spring qualities.

The next larger X-L filament tube is the UV-204-A radiotron, shown in Figure 10. This tube has an output rating of 250 watts. The normal plate voltage is 2,000 and the filament requires 11 volts and 3.85 amperes or 42.5 watts. The maximum plate power dissipation is 250 watts.

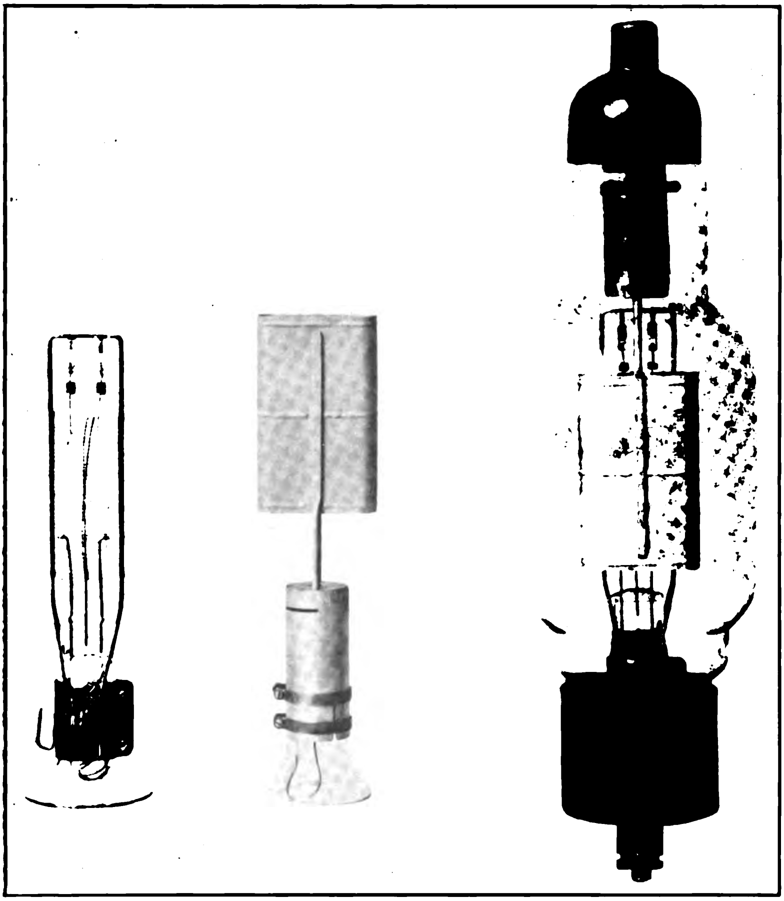

FIGURE 10—Radiotron UV-204-A 250-Watt Output

As already stated, this tube directly supersedes the UV-204, and on account of the interchangeability requirements the plate characteristics are almost identical with those of the UV-204. Consequently, the later tube may be used in sets designed for the UV-204, the only change required being an increase in the rheostat resistance to allow for lower filament current.

The usual static characteristics of the UV-204-A tube are given in Figures 11 and 12. The amplification constant averages approximately 24 and the internal plate resistance at 2,000 volts, and zero grid is approximately 4,700 ohms, giving a mutual conductance, under these conditions, of 5,100 micromhos.

The total electron emission at rated voltage is approximately 5 amperes, which is much higher than the emission of the UV-204

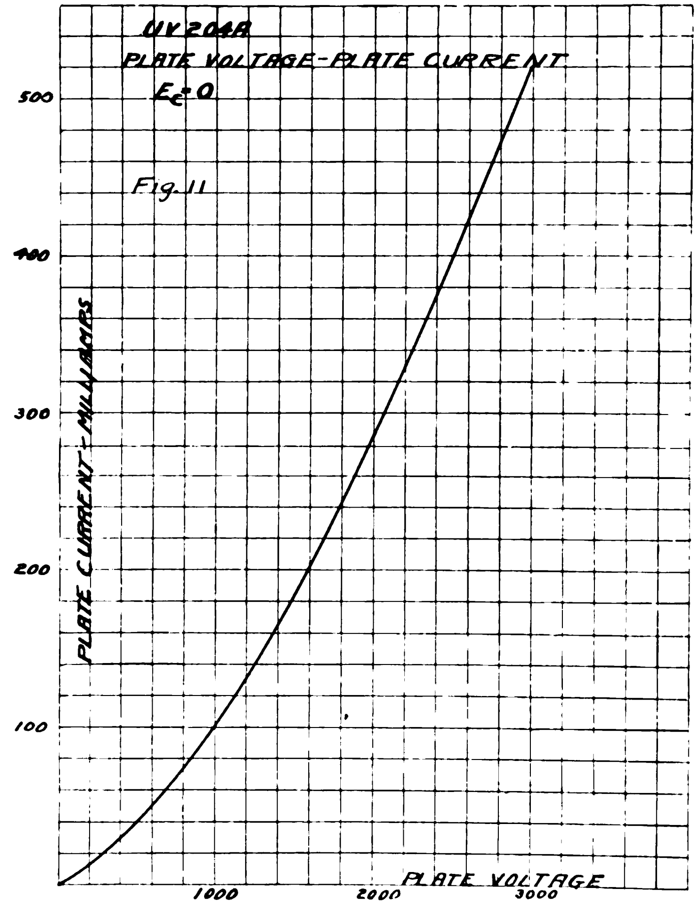

FIGURE 11

filament. This results in somewhat better performance in telephone circuits where the peak values of current are likely to be considerable.

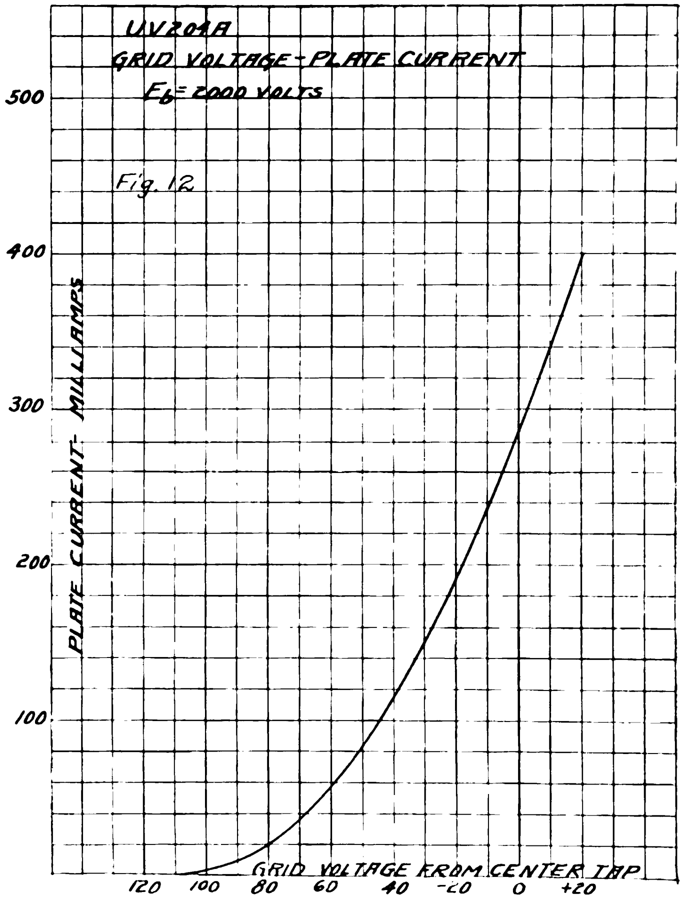

FIGURE 12

For outputs greater than 250 watts, it is quite possible to connect as many as four 250-watt tubes in parallel. However, there are always certain difficulties present in parallel operation of more than two or three tubes, and to supply the demand for a low voltage, high output tube, the UV-851 radiotron has been developed. At present this is the largest power tube utilizing the X-L filament. There is already another one-kilowatt radiotron, the UV-206, but this contains a pure tungsten filament and normally requires a plate voltage of 10,000 volts or higher.

The normal plate voltage of the UV-851 is 2,000 volts, and the maximum plate power dissipation is 750 watts when the tube is used as an oscillator. Under these conditions, the tube is capable of giving a radio frequency output of one kilowatt or more. The filament requires 15.5 amperes at 11 volts, or a power consumption of 170 watts. The total electron emission is about 20 amperes.

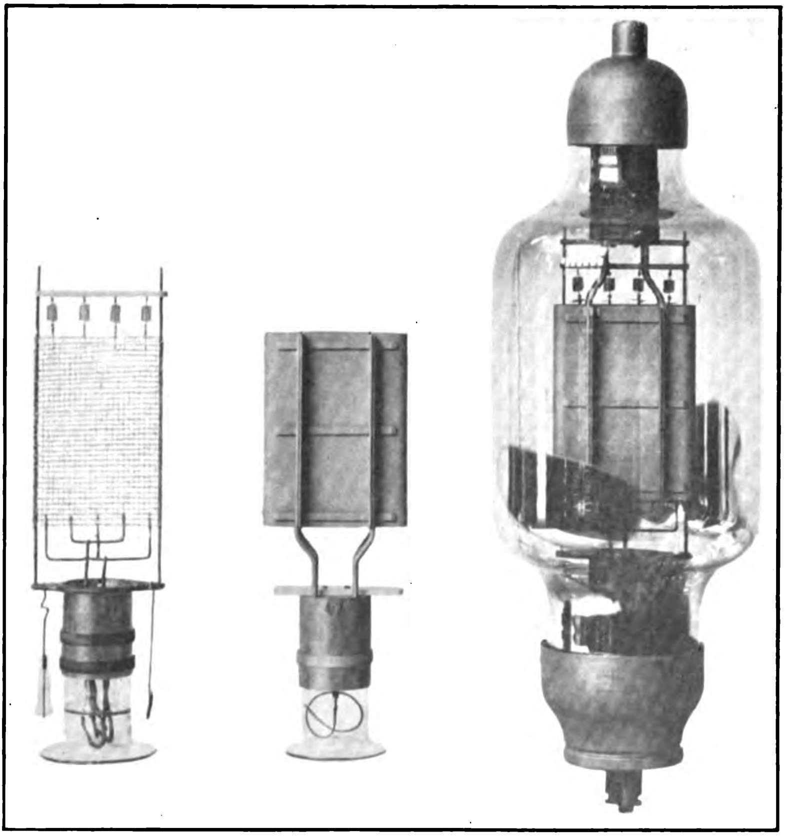

FIGURE 13—Radiotron UV-851 1 KW Output

Figure 13 shows the external appearance of the tube and the anode, cathode and grid structures.

The X-L filament, while bringing about marked improvements in all of the tubes in which it has superseded the pure tungsten filament, has been of even greater importance in the design of the UV-851. A tube of this type and size would have been almost impossible to build with a pure tungsten filament. This

is evident when it is realized that a pure tungsten filament would have required at least 600 to 700 watts to obtain the necessary electron emission. This power would have been dissipated almost entirely inside the plate and would have resulted in considerable over-heating when added to the normal dissipation due to internal plate circuit losses. Furthermore, the plate char-

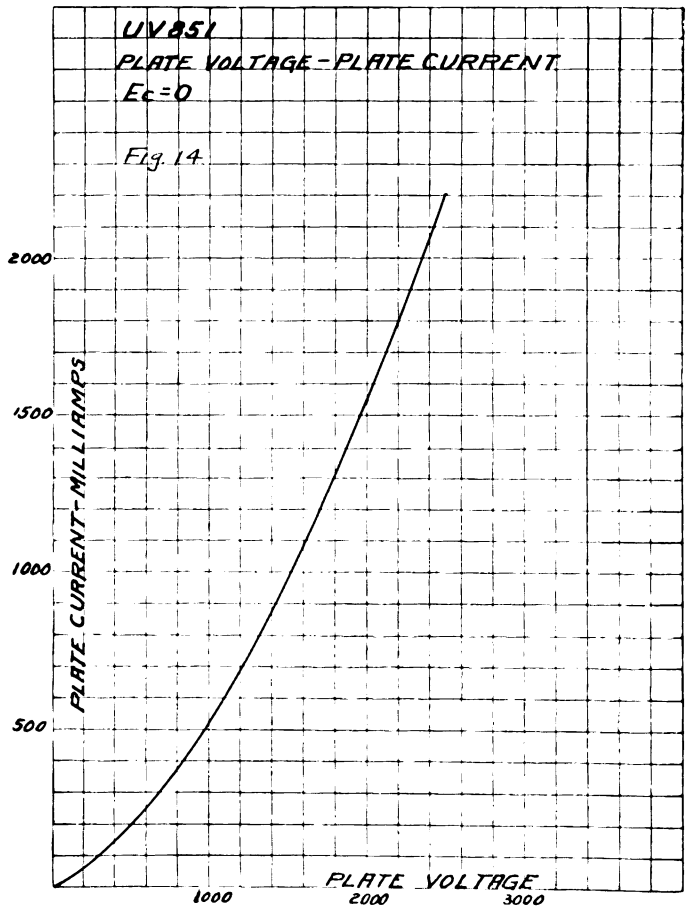

FIGURE 14

acteristics required to give one-kilowatt output with reasonable efficiency at the relatively low plate voltage of 2,000 volts would have been unattainable with a tungsten filament.

The amplification constant of the UV-851 is approximately 20 and with 2,000 volts on the plate and zero grid the internal plate resistance is approximately 850 ohms and the mutual con-

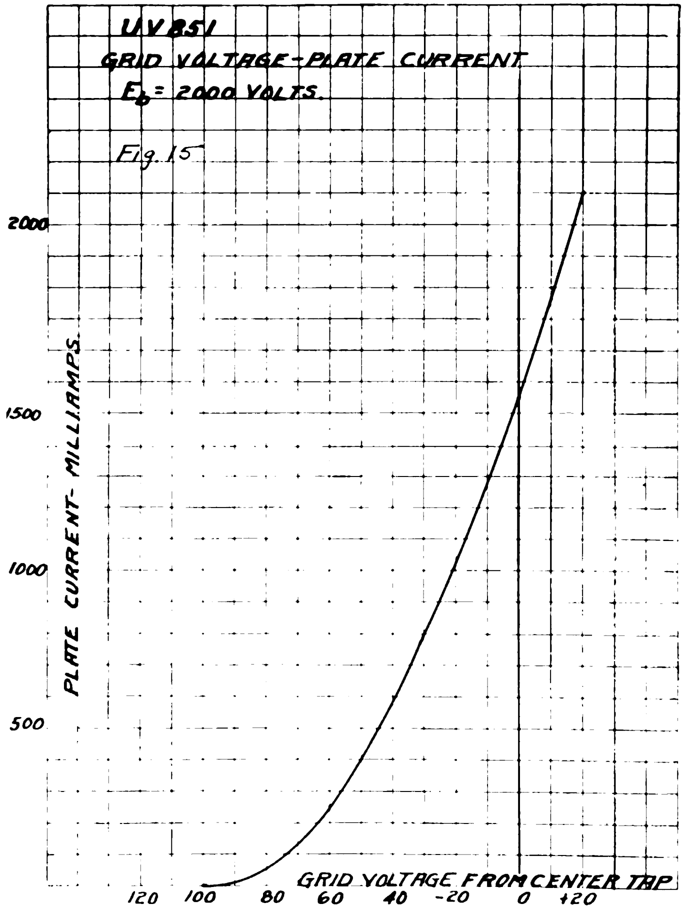

FIGURE 15

ductance, therefore, is 23.5 milliamperes per volt or 23,500 micromhos.

The static characteristics of the tube are shown in Figures 14 and 15.

The mechanical construction is of some interest on account of the large size of anode and grid structures and also due to the fact that four parallel filaments are used. The advantage of the parallel type of filament has already been explained. It is particularly important in a tube of this size, because of the necessity of reducing as much as possible the plate losses inside the tube. The grid also is somewhat different from that ordinarily used, being made of a heavy square mesh of molybdenum wire.

The grid frame is anchored to the plate structure with an insulator so as to maintain the proper position of the grid inside the anode. The plate is of exceptionally heavy construction and its surface is so treated as to give good radiating properties. Proper filament tension is maintained, as in the UV-203-A and UV-204-A, by helical springs, which are protected by small metal discs from excessive heat.

The tubes described above, while not including all that have been designed with X-L filaments, serve to illustrate the most important advantages of this type of filament over the older pure tungsten type, resulting in improved operating characteristics in tubes of such sizes as were already in existence, and permitting the design of a new class of tube, the manufacture of which otherwise would have been impractical if not impossible.

For reference a table of the electrical constants of the X-L filament tubes is given in Figure 16.

Radiotrons	UV-210	UV-203-A	UV-211	UV-204-A	UV-851
Rated Output (Watts)	7.5	50	50	250	1,000
Max. Plate Dissipation	15	100	100	200	750
Filament Voltage	7.5	10	10	11	11
Filament Amperes	1.25	3.25	3.25	3.85	15.5
Plate Voltage	350	1,000	1,000	2,000	2,000
Plate Current Oscillating (m.a.)	60	125	125	200	875
Amp. Constant (Approx.)	7.5	25	12	24	20
Plate Impedance (Approx. ohms) at zero grid and rated plate voltage (See Note)	3,500	5,000	1,900	4,700	850
Mutual Conductance (Micromhos) at zero grid and rated plate voltage (See Note)	2,150	5,000	6,300	5,100	23,500
Plate Current (m.a.) at zero grid and rated plate voltage (See Note)	70	120	320	275	1,550
Dimensions Over-all (inches)	5¼	7⅞	7⅞	14¼	17½

NOTE: These figures are given for comparison only and do not necessarily apply to all conditions of normal operation.

FIGURE 16—Characteristics of X-L Filament Transmitting Radiotrons

SUMMARY: The properties of the X-L or thoriated tungsten filament are discussed with particular reference to the suitability of this material for use in power tubes and its advantages over other materials. Comparisons are given between pure tungsten and thoriated tungsten filaments in electron emission characteristics and effect on tube design and performance.

Several power tubes containing X-L filaments are described in detail. The improvements due to the use of the X-L filament are illustrated by comparison of these tubes with older types of tubes containing pure tungsten filaments.

DETECTING CHARACTERISTICS OF ELECTRON TUBES*

By

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For a number of years the electron tube has been regarded as the nearest approach to an ideal detector of radio power that has yet been devised. Its chief advantage over the simple contact rectifier lies in its stability and ease of control, and in its amplifying property, particularly that factor which permits the building up of the input voltage by means of regeneration. In fact, this regenerative faculty has provided such a simple and easily controllable means for increasing the detector efficiency that the minor factors controlling the performance of the tube as a detector have been rather overlooked, and very little attempt has been made to develop the possibilities of controlling these minor factors in such a way as to increase the efficiency of the tube as a non-regenerative detector. However, the well-known disadvantages of the use of regeneration in its simplest form are becoming of greater and greater importance with the continually increasing number of listeners in radio; and the use of the tube as a detector is being more and more restricted to the simple rectifying function.

This being the condition now in existence, it becomes of more importance than ever before to examine carefully the operation of the tube as a non-regenerative detector; and to determine more thoroly than has yet been done the importance of the various characteristics which enter into the operation of the tube as a detector, and the factors which control these characteristics. Up to the present time the practice among the manufacturers has been to work toward a general purpose tube, on the theory that almost any kind of a tube is good enough as a detector if the characteristics which determine its efficiency as an amplifier are sufficiently emphasized. With the development of the art, the advantage of using one type of tube for all purposes is decreasing

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in importance, and an increasing appreciation of the large part played in set performance by the efficiency of the detector is bringing out to an ever increasing degree the need of emphasizing the characteristics of a tube which are of primary effect in ts performance as a detector.

The present paper is a report of some preliminary studies of the tubes now in existence, with the object of opening up the field for the development and control of the operational factors of the detector tube; and while the work has not yet progressed far enough to venture any predictions as to what may eventually develop along these lines, it offers some interest as evidence of the improvement that is possible even in existing tubes over the commonly used conditions of operation.

As a preliminary to developing criteria which will enable the performance of a tube as a detector to be easily and accurately predicted and as a check on the proposed method of measurement, a study was made of a normal tube of a type much used at the present time for either detector or amplifier. In this particular tube, the filament is designed to operare at 1.1 volts and 0.25 amperes for general service, and it has a plate impedance of 14,000 ohms, an amplification factor of 5.6, and a mutual conductance of 400 microamperes per volt when measured with a plate voltage of 45 volts with the grid tied to the negative end of the filament. Following the well-known development for the performance of a tube as a detector when used in the circuit shown in Figure 1, the expression for the change in plate current produced by impressing a signal voltage E in the grid circuit is as follows:[1]

$$\Delta I_P = \frac{E^2}{2} \frac{\frac{d^2 I_G}{d E_G^2}}{\frac{d I_G}{d E_G}} \frac{d I_P}{d E_G} \qquad (1)$$

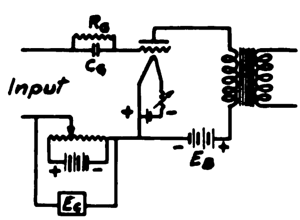

FIGURE 1—Circuit Used for Detector Test

[1] J. H. Morecroft, "Principles of Radio Communication," page 456.

From this expression it appears that the output obtained from a given incoming signal is proportional to the product of the rate of change of the slope of the grid characteristic by the slope of the mutual characteristic, and inversely proportional to the slope of the grid characteristic.

From an analysis[2] of the action taking place in a detector circuit comprising a grid condenser and grid leak, it is at once apparent that certain conditions must be met in order to obtain efficient operation, before the characteristics of the tube itself become effective. The condenser must be of such a value that it will be fully charged up to the maximum voltage of the incoming signal during the time in which the signal is impressed. For example, if we assume a signal frequency of 1,000,000 cycles per second, modulated at a frequency of 10,000 cycles (which is well within the conditions of ordinary telephony) we find that a tube having a grid conductance of 5 microamperes per volt will allow a condenser of 0.00025 microfarad capacity to become charged up to the peak voltage in one half the time of a complete cycle of modulation. Such a condenser will require a leak of 400,000 ohms in order to discharge completely in the same length of time. This rough estimate gives us semi-arbitrary values of grid condenser and leak which should give good operation with a normal tube of the type under discussion, and these values are used for the purpose of this experimental check.

Confining the experiment for the present to conditions such as are often met in practice, characteristic curves have been drawn for plate voltages of 10, 20, 30, and 40; the filament being maintained constant at 1.0 volt. In drawing these curves, the grid return was connected in each case to the positive end of the filament, making a constant positive potential of 1.0 volt on the grid referred to the negative end of the filament.

Figure 2 shows the variation of grid current with grid voltage for the values of plate voltage shown above. Figure 3 shows variation of plate current with grid voltage over the same range. These curves were obtained experimentally in the customary way, and were used as the basis for deriving curves from which the values of the factors in the expression for output could be obtained.

The first of these derived curves shows the variation with grid voltage of the slope of the grid current-grid voltage curve. This value, $\frac{dI_G}{dE_G}$, is plotted against E_G in Figure 4. The initial

[2] J. H. Morecroft, "Principles of Radio Communication," page 461.

conditions under which the experiment was carried out demanded a value of $\frac{dI_G}{dE_G}$ of at least 5 microamperes per volt. Consequently, further discussion is confined to those portions of the curves above the horizontal line representing this limiting value.

Figure 5 shows the values of the second derivative, $\frac{d^2 I_G}{dE_G{}^2}$, as as obtained from the curves of Figure 4. Figure 6 shows the curves of $\frac{dI_P}{dE_G}$ plotted against E_G, as derived from the mutual characteristic curves of Figure 3.

From the three derived curves of Figures 4, 5, and 6 the values of ΔI_P for various settings of E_G are calculated from the equation given above; and the resulting curves are plotted in

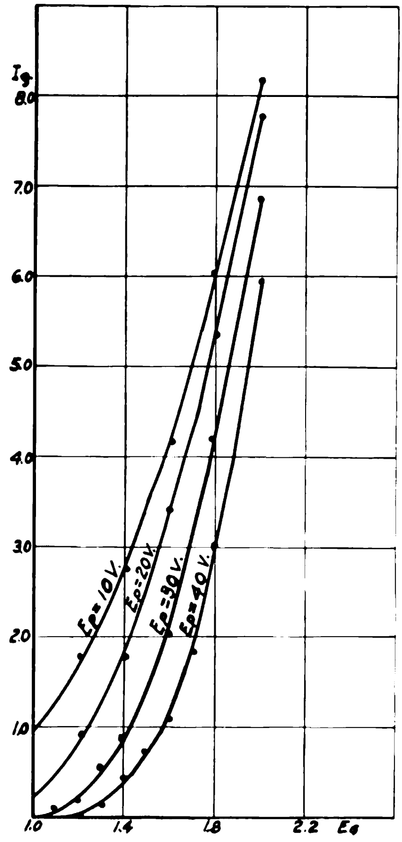

FIGURE 2—Tube DT-1, Grid Characteristics

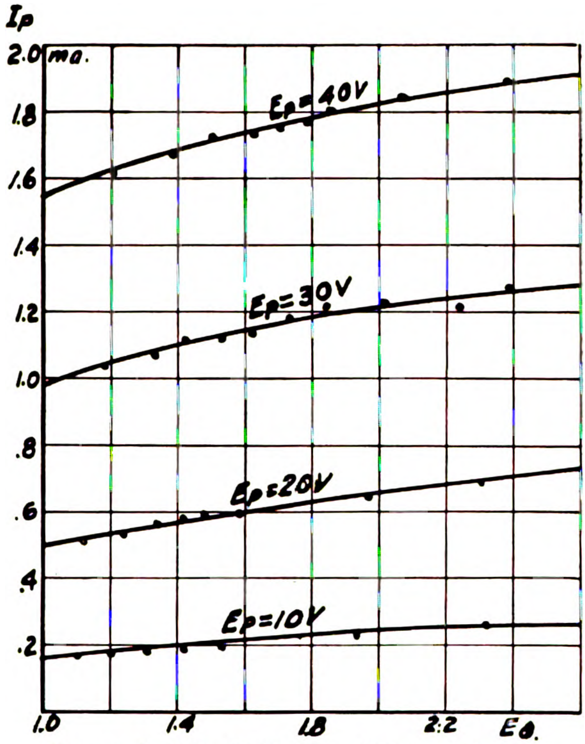

FIGURE 3—Tube DT-1, Mutual Characteristics

Figure 7. All of these curves show maximum signal strength at the lower limit of the region explored. The sharp falling off of signal strength, as $\dfrac{dI_G}{dE_G}$ increases because of increased E_G, shows plainly the danger in using too high a positive voltage on the grid.

In order to check experimentally the findings predicted from theoretical considerations and plotted in Figure 7, the strength of the output obtained from a standard signal voltage impressed in the circuit of Figure 1 was measured for a number of values of grid voltage and plate voltage within the region covered by the curves of Figure 7. The arrangement of circuits used is diagrammed in Figure 8. The source supplying a modulated radio frequency signal of constant strength was coupled loosely to the dummy antenna used in connection with a tuner and the circuit of Figure 1 in the detector D. The output of the detector was amplified by a shielded two-stage audio frequency amplifier A, and the peak voltage on the secondary terminals of a transformer in the plate circuit of the last tube was measured by means

of an electron tube voltmeter of the type described by R. A. Heising.[3] This circuit (Figure 9) is used as follows: The tube is set at a certain grid bias sufficient to bring the plate current at normal filament operation, nearly to zero. This bias is determined by the point on the potentiometer P to which the positive terminal of direct current voltmeter V is connected.

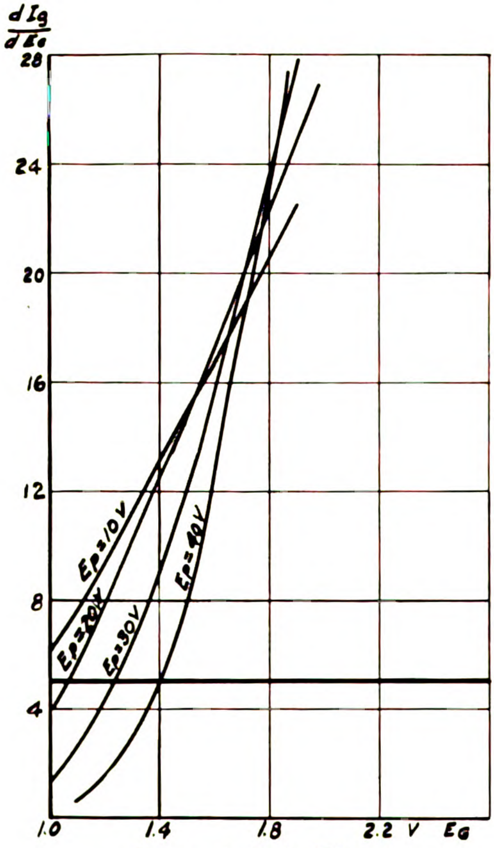

FIGURE 4—Tube DT-1

With no signal applied to the terminals of the transformer T, the filament is adjusted until the microammeter M shows a definite very small reading. When an alternating voltage is applied to the terminals of T, a reading is obtained on M because of the rectifying property of the tube as operated at the lower end of its mutual characteristic. The slider on P is now moved in such a way as to make the grid more negative, until the read-

[3] J. H. van der Bijl, "The Thermionic Vacuum Tube," page 367.

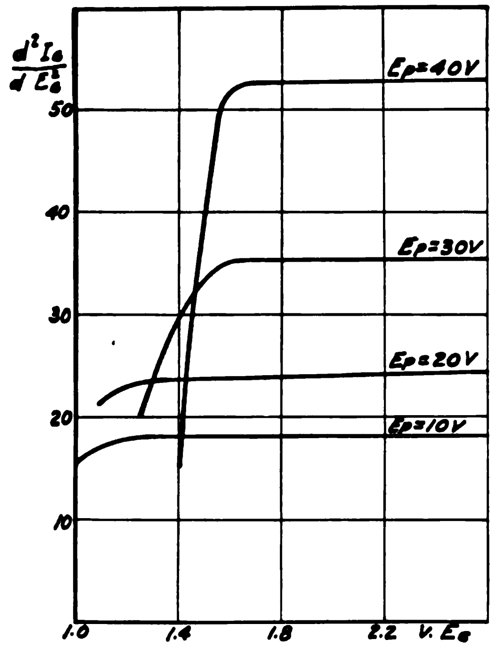

FIGURE 5—Tube DT-1

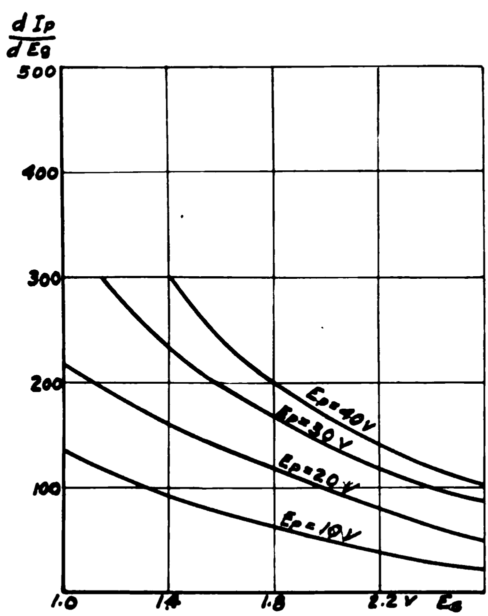

FIGURE 6—Tube DT-1

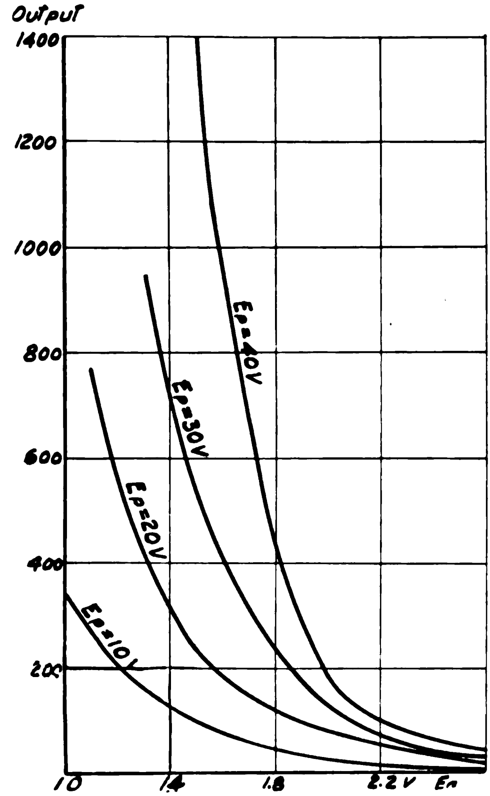

FIGURE 7—Tube DT-1, Derived Curves. Detector Output

ing of M is the same as it was when no signal was coming in. The amount of additional grid bias necessary to bring about this result, as measured on the voltmeter V, is equal to the peak voltage of the signal in the secondary of the transformer T, and, therefore, proportional to the change in plate current in the detector tube, within reasonable limits of operation.

It has always been a matter of extreme difficulty to obtain satisfactory measurements of this kind, particularly in a building subject to disturbances from other radio sets, electric motors, and other sources of electrical disturbance. This difficulty was overcome to a considerable degree in the present case by performing the whole experiment inside of the cage shown in Figure 10. This cage is made of expanded metal screen, having a diamond mesh of approximately 1 inch by 1½ inches, the metal strands being roughly the size of number 8 gauge wire. All

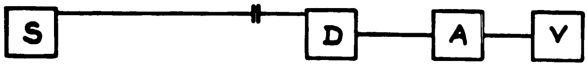

FIGURE 8—Diagram Showing Arrangement of Apparatus for Testing Detector Output

seams were carefully covered with metal straps bolted to the supporting steel framework of the cage, and the door was supplied on all its edges with copper lining strips which formed a close electrical contact with the door frame when the door was closed. The cage was thus completely a series of closed circuits, and proved very effective in screening out radio frequency disturbance of all kinds. For example, a receiving set comprising two stages of tuned radio frequency amplification, detector, and two stages of audio amplification gave an interference output that was barely audible in head phones inside of the cage, altho outside in the room the noise from commutators and other electrical equipment in the building furnished a continual roar in the loud speaker of such volume as to make it a matter of extreme difficulty to talk against it. Station KDKA, at a distance of approximately one-quarter of a mile, could be heard with the head telephones on the same set inside of the cage, but was easily eliminated by tuning. The cage was insulated from the floor (which is of a highly conducting material) and there were no leads running into the cage from the outside.

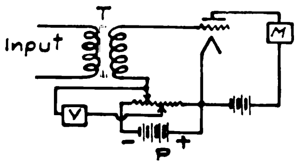

FIGURE 9—Circuit of Tube Voltmeter

By the use of this cage, and by observing all possible precautions to insure the constancy of operation of the component parts of the set-up, particularly of the source of signal and of the shielded amplifier *A*, it was found possible to obtain easily reproducible results in the determinations of the relative output obtained for various operating conditions of the detector tube *D*.

A correction which it was found necessary to make in assigning values of the grid voltage E_G occurred because of the change in resistance of the grid leak with the amount of current passing thru it. The reason for this change is not apparent, but since it seemed to be constant and reproducible it was merely taken

account of as a correction in determining the value of grid battery E_C necessary to provide the desired grid voltage E_G.

The curves of Figure 11 show the variation of signal output with voltage of grid battery, and are of the same general form as the curves of Figure 7, derived from the static characteristics of the tube. As a check on the closeness of agreement, the curves

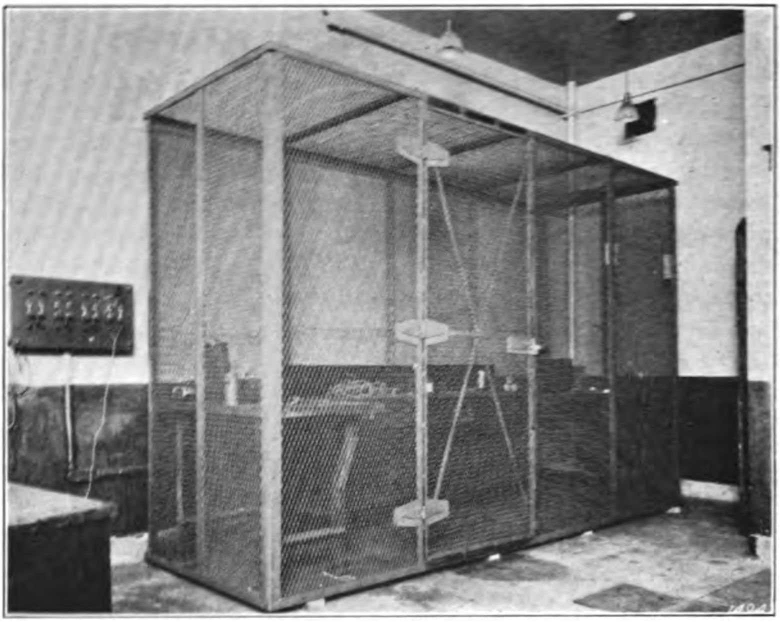

FIGURE 10

of Figure 7 were re-drawn, assigning the arbitrary value of 100 units to the maximum point shown and calculating other points on the curves on this basis. Translating the value of grid battery in the curve, Figure 11, into the corresponding voltages on the grid by subtracting the drop due to the resistance of the grid leak and referring to the maximum reading as 100 units, the experimental values of output are plotted as crosses on the same sheet. Figure 12 shows the closeness of agreement between the calculated curves and the experimental points.

It is apparent that this method of measuring the output from the detector tube gives results which agree satisfactorily with the theoretical values obtained from the static characteristics, and therefore furnishes a quite accurate means of making quick determinations of the effect upon detector efficiency of changing the operating conditions of the tube.

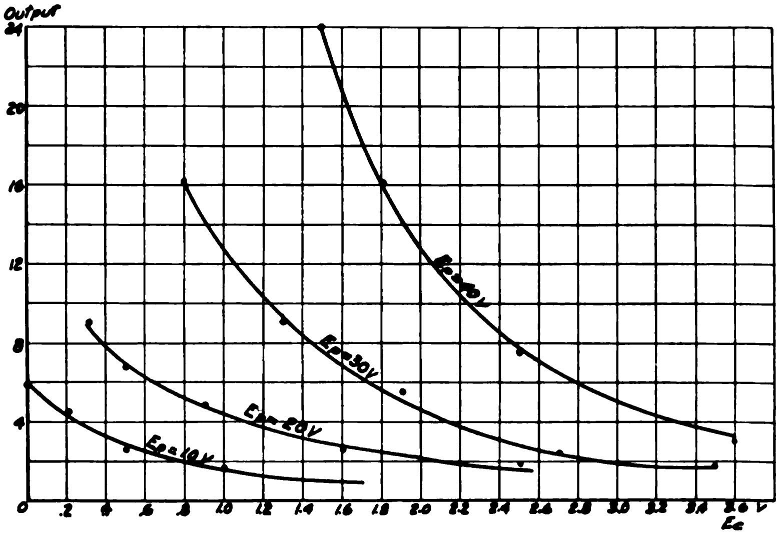

FIGURE 11—Tube DT-1, Experimental Curves, Detector Output

The curve of Figure 13 shows the results obtained with tubes of three well-known commercial types, the grid return being to the negative end of the filament in each case. In this test the

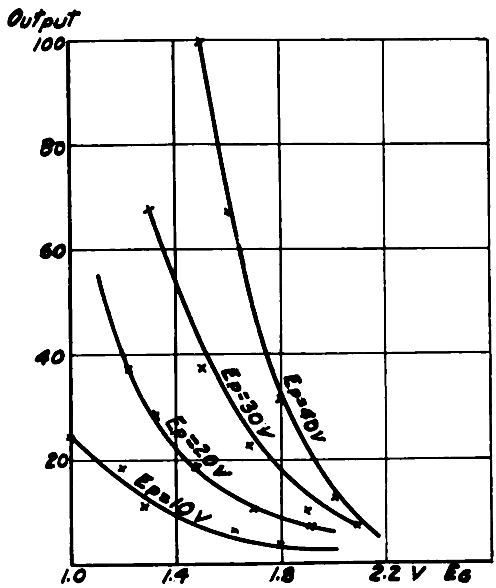

FIGURE 12—Tube DT-1, Comparison Experimental and Derived Curve. Detector Output

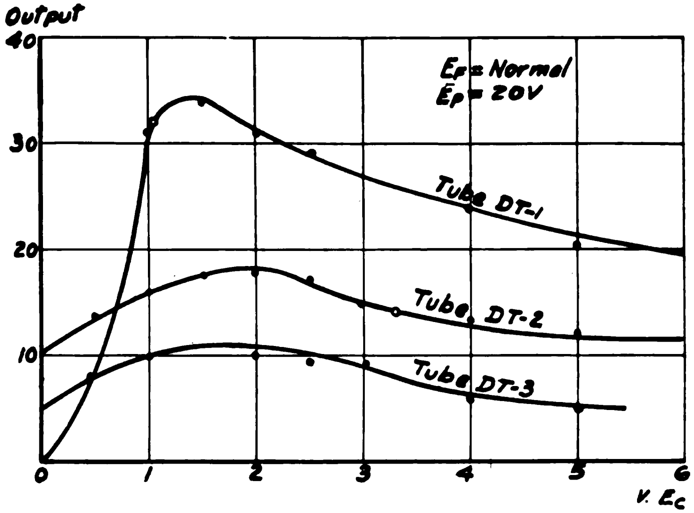

FIGURE 13—Variation of Detector Output with Grid Voltage

grid condenser and leak were of values customarily used, and the filaments were operated at rated supply voltage. The plate voltage was 22½ volts. The circles on the curves indicate the value of filament voltage drop and show what may be expected of these tubes when operated with zero grid bias but connected to the positive end of the filament. In the case of these individual

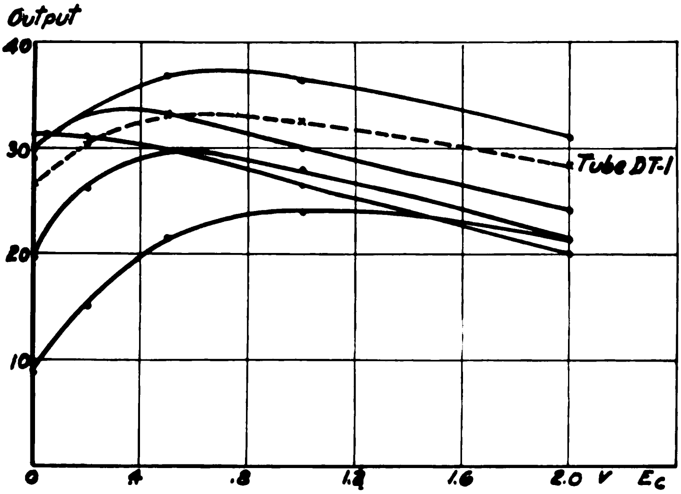

FIGURE 14—Output from Various Tubes of 1 Type
E_f = Normal E_P = 20 V.
Grid to Positive Filament

tubes it appears that the one having the 1.1-volt filament happens to approach the required grid bias conditions most closely when connected in this way.

The curves of Figure 14 were drawn for six tubes of this type with the grid connected to the positive end of the filament, normal filament voltage and a plate supply of 22½ volts being used. Altho these tubes showed approximately the same values of the plate and mutual characteristics ordinarily used to describe tube performance, their operation as detectors showed a wide range of values under any single fixed operating condition. It will be noted that all of these tubes show the normal operation for this type as given in the curve of Figure 13, at some particular operating point, but that in some cases the proper grid bias is quite different from that obtained in the customary connection in the set. This result explains many of the phenomena that have been observed regarding the improvement which may be obtained in the operation of a set by juggling the tubes around and emphasizes the importance of the problem outlined in this paper.

Research Laboratory, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania.

January 19, 1925.

SUMMARY: It is pointed out that owing to the progress of the radio art the opinions heretofore held as to the importance of the part played by detector efficiency in a receiving set are in need of revision.

Taking the well-known analysis of the operation of a detector tube with condenser and grid leak, curves are derived from the static characteristics of a typical general purpose tube showing the performance of the tube as a detector under certain conditions of operation.

A method is described of measuring the output of a detector tube with a standard incoming signal, and experimental results obtained with the tube used for deriving the curves are compared with those obtained from theoretical considerations, showing that the method can be used to give a true picture of the effect on detector efficiency of variations in operating conditions.

Sample curves are given, showing the wide variations obtained in the efficiency of certain types of standard tubes by relatively slight changes from the customary operating conditions, and also the variations in efficiency of a number of similar tubes under normal operating conditions.

LIFE TESTING OF TUNGSTEN FILAMENT TRIODES*

By

William C. White

(Research Laboratory, General Electric Company, Schenectady, New York)

General

Very little has been published on triode life, a most important subject, and probably for this reason very few of those interested in radio have any reliable information regarding it.

The most common opinion is that the life of a certain type or make of triode is a sort of constant, like the amplification constant. This viewpoint is entirely incorrect and it is the object of this paper to show what a variable factor triode life really is.

It is not uncommon to hear of a life test being made on a few triodes of one type under conditions of doubtful constancy and the result used to indicate the life of that type under all conditions over a long period of time.

This subject has in many ways marked similarities to the study of vital statistics in the field of medicine.

The gathering, compiling, and studying of vital statistics is not made so that any individual may estimate how long he is going to live, but in order to furnish information to help prevent or combat the various diseases and causes of death.

In a similar way the manufacturer of triodes obtains great advantage in observing the behavior during life of his product and length of its life with operating conditions accurately controlled.

A great deal of data is continually being published on the constants and characteristics of triodes. In practically all of these cases the triodes are comparatively new. The user is primarily interested in their performance thru at least several hundred hours of use. Therefore, initial constants and characteristics are only a small part of the performance rating of the tube.

Without going thoroly into the subject, one might think that

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it is a simple matter to set up and operate a triode life test, but experience has shown that unless great care is continually taken on many details the results are worse than useless, because they may be so misleading.

Some of these details are such things as, what constitutes end of life, electrical conditions of operation so results are comparable to practice, and obtaining accurately the desired information on a large number of tubes with a reasonable cost.

The subject is a complex one because the effect of some particular variation of design or manufacturing process may give a resultant change in life on one test, whereas, the same test conducted a year later, may give an opposite result, due to still another slight change which has occurred in the meantime.

To be of real value, life testing must be kept up continuously with triode production. In general, to check up a regularly manufactured product, it has been found more satisfactory to put a few samples on life test at regular brief intervals, rather than relatively larger quantities at less frequent intervals.

Even with great care in all respects some life test results are difficult to understand or analyze.

Reasons for Life Test

The most important reasons for the life testing of triodes are as follows:

(1) To life test new types before their regular use has begun, so as to be certain that no details have been overlooked which would prevent the realization of a satisfactory operating life.

(2) To enable the best choice to be made between several different proposed details of design. As will be pointed out later, very often some little, apparently insignificant, detail will have a pronounced effect upon life.

(3) To detect marked changes in the life quality of the triodes as regularly produced.

(4) To aid in the choice of the best raw materials conducive to long life.

(5) To determine the most satisfactory exhaust method.

(6) To learn if any of the characteristics change during the operating life and, if so, to what extent.

(7) To learn the effect on life of different combinations of operating conditions such as:

(a) position of mounting
(b) room temperature
(c) filament voltage

(d) plate voltage
(e) plate current
(f) grid bias voltage
(g) intermittent operation

Apparatus Used

The apparatus to be described is located in the Research Laboratory of the General Electric Company at Schenectady, New York.

(1) *For Receiving Triodes*

The general arrangement of life testing racks for receiving triodes is shown in Figure 1. Each of these large racks is divided into a number of smaller sections and on each of these sections the filament, plate, and grid voltages can be adjusted to any desired value within certain ranges. This is accomplished by means of small control panels, as shown in Figure 2, which is a close-up view of four of the sections.

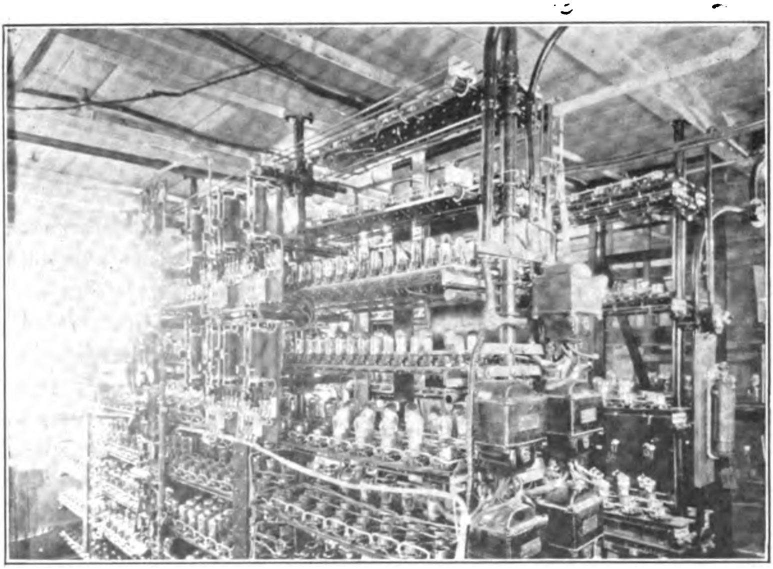


Figure 1

Plate voltage is supplied at 125 volts from the building direct current power lines, as it has been found that the regulation of these lines is close enough so as not to affect the results. This voltage is adjusted by means of a slide wire rheostat in a potentiometer connection, and the plate voltage on the triodes is indicated by a small voltmeter.

The grid voltage is supplied from a small low voltage direct current generator, and it is adjusted by a small circular potentiometer rheostat, and the actual bias voltage indicated by a small voltmeter.

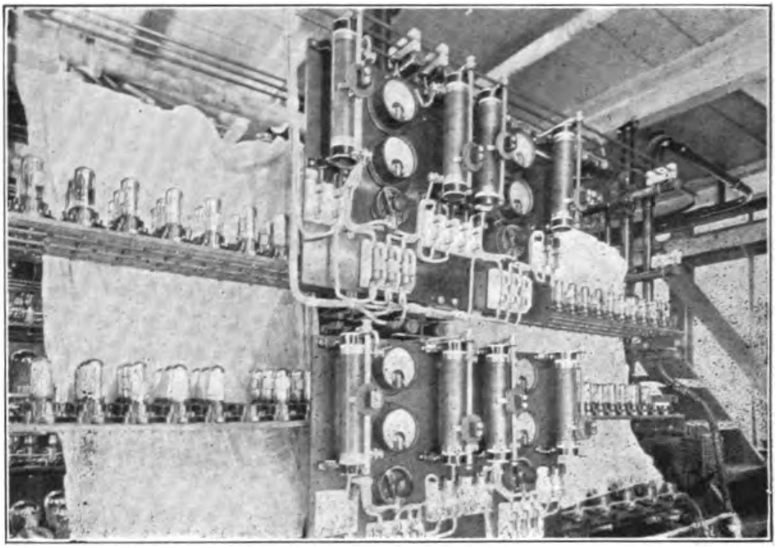


FIGURE 2

The filament voltage is supplied thru special motor generator sets equipped with voltage regulating devices so that the voltage is held well within plus or minus one percent. Owing to the fact that many different voltages are employed, alternating current has been found most practical. This low voltage alternating current is supplied by a transformer on each small section and controlled by a third slide wire resistance. The filament voltage is measured by a high grade and frequently calibrated portable alternating current voltmeter, especially designed to require less current for operation than the usual instruments. When making a filament voltage measurement, leads from this voltmeter are clipped to the filament terminals of one of the sockets of the section.

It is true, of course, that practically no receiving triodes are actually operated on alternating current. We know from experience, however, that there is no very great difference between alternating current and direct current for filament operation, and, as these life tests are conducted from the viewpoint of obtaining comparative results, rather than results to show the highest

possible life figures, it is believed that alternating current is preferable for filament operation in life tests where a great number of different values of filament voltage must be used.

The main generator for supplying the filaments delivers power at a thousand volts and this is stepped down to approximately 115 volts for the small section transformers behind the main life test distributing panel. The individual transformers for supplying the small sections are located so that the low voltage leads are only a few inches long. By this method of distribution, no heavy currents are carried thru any great distance and voltage variation, due to change of load, is minimized. Also, variations of load on one section do not appreciably affect the voltage on other sections.

A large number of the triodes life tested are run under certain standardized conditions, and it might seem best, therefore, to have one large section devoted to each type. It has been found, however, that sections containing a very large number of triodes give a great deal of trouble from high frequency oscillations set up in the leads and wires connected with the triodes. These oscillations may seriously change the bias voltage conditions, and even have been found in some cases to vary the filament voltage considerably. They are often of such an extremely high frequency that it is difficult to prevent their occurrence in all parts of a section, and for this reason experience would indicate that sections containing not more than fifty to one hundred tubes are preferable.

(2) *For Transmitting Triodes*

A view of one portion of the transmitting rack devoted to the small power triodes is shown in Figure 3. A wire screen protective covering is used over the racks and at intervals gates are provided to gain access to the triodes. This protection is advisable, as the plate voltages used are from 500 to 2,000. The gates are provided with contacts so that this high voltage is automatically disconnected when a gate is opened.

Transmitting triodes are, in general, tested non-oscillating; that is, they are tested at rated filament and plate voltage with the plate current adjusted to give a certain power dissipation from the plate by means of a proper grid bias voltage. This condition of operation is similar to that obtained in the use of the triode as a modulator, or power amplifier, when no alternating voltage for amplification is applied to the grid.

Because of the high plate voltages used, three fixed values of

direct current plate voltage are supplied to this life testing rack. These are 2,000, 1,000, and 500 volts. A number of separate sections, however, are available for each type of triode and these sections are provided with independent control of filament voltage and grid bias voltage.

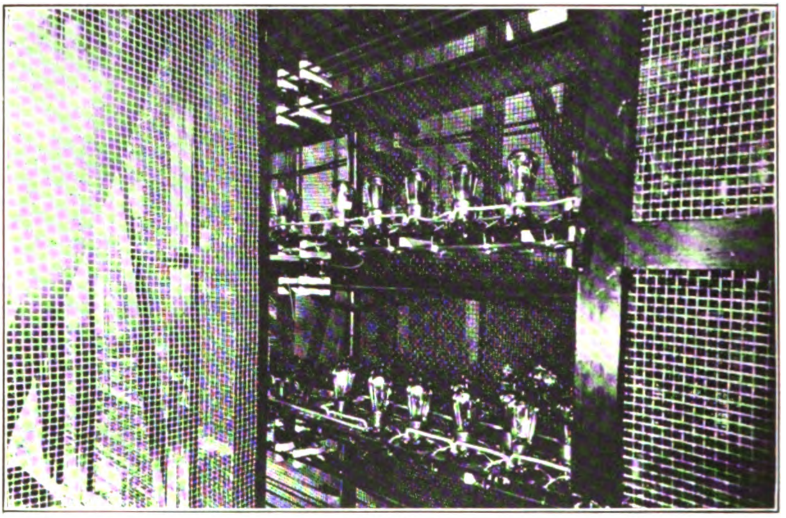


FIGURE 3

In the case of transmitting triodes, alternating current is also used for filament operation, but this is entirely logical as most of the power triodes are used in this way.

Great care must also be taken to prevent very high frequency oscillations in these life testing sections. In general, a high resistance in the grid lead of each tube is employed, and the number of tubes per section is also limited.

The power triode rack, which has just been described, is used for testing triodes of the five-watt to the one-kilowatt, 2,000 volt, types inclusive.

Some life tests are also run on the high voltage higher power triodes, but in this case the numbers tested are so small relatively and the tests are of such a special nature, that they will not be included in this paper.

(3) *Miscellaneous*

Relays are provided, operated from the grid bias voltage, so that in case this voltage fails, the filament voltage or plate voltage, as is most convenient in any particular case, is discon-

nected from the triodes. This is a necessary precaution because many types of tubes would be ruined very quickly by overloading if the grid bias voltage failed for any considerable length of time.

A recording voltmeter, with charts changed daily, is used in connection with the main filament voltage generator. This instrument shows the voltage regulation attained, and indicates any failure of this supply for even a short period of time. During the night hours the life test runs with only such attention as can be given it by a night watchman making his rounds about once an hour, so that these features are necessary.

Figure 4 shows an electrically driven commutator having two large drums. One of these drums makes a revolution once every twenty-four hours and the other once every hour. On the

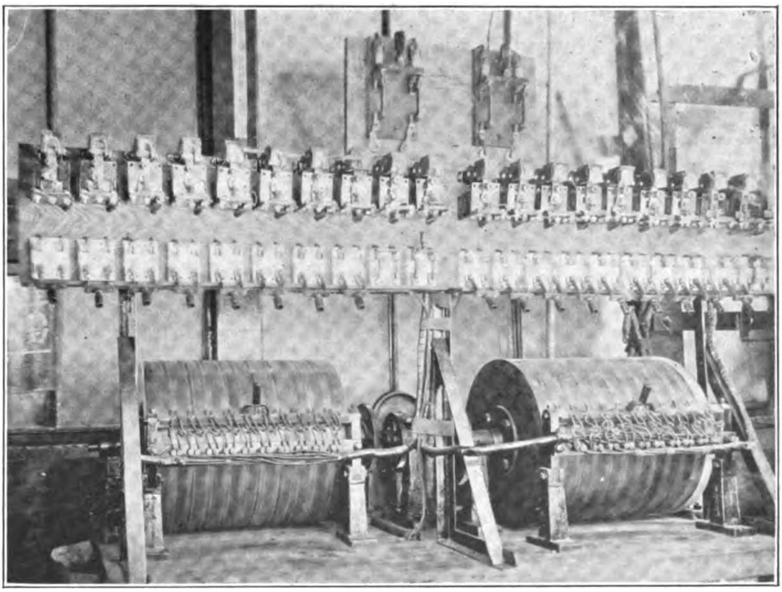


FIGURE 4

surface of these wooden drums can be easily fastened thin strips of copper which can be attached in such sequence as to give any desired time control. Pairs of light brushes press against these copper strips on the drums and in circuit with them are included relays so that plate, grid, and filament voltages on any section of the life test can be thrown on or off or varied by certain definite steps on any desired time schedule.

Method of Tests

All triodes to be life tested are labeled and each lot is given a designation, which is coded to some extent, to indicate briefly the identity and classification of the triode.

For each lot of triodes to be tested a "work sheet" is made out, giving all the information about the identity of the triodes, such as the purpose for which the test is being run; the conditions under which these triodes are to be life tested; the readings to be taken initially and during life, and the person or persons to whom the results are to be reported. When the life testing has been completed, a summary form is filled in, giving as briefly as possible the result and desired information. An extra copy of this summary is filed under the type of triode tested, so that all life information on a certain type of tube can be found in one place.

Usually, the electrical conditions of test are chosen to represent at least the most severe conditions given by the rating of or instruction sheets accompanying the triode. Very often the conditions are made even more severe than this. This is done, first, to give a factor of safety beyond normal operating conditions and, second, to give the results desired as quickly as possible. Again, it should be pointed out that the important reason for conducting life tests is to obtain relative values and the actual figure obtained for a given lot of tubes taken by itself is not usually of great interest or value.

From time to time these life test conditions for each type of triode are changed to take account of changing radio practices. In so far as possible, the number and frequency of these changes are kept to a minimum so that the results obtained on a given type of triode are comparable over a considerable period of time.

In general, for each type of triode, a standardized schedule is followed, giving the hours at which triodes are removed from life test to be given electrical characteristic tests. Such a schedule generally includes electrical characteristic tests initially and as an example at the end of 25, 50, 100, 250, 500, 750, 1,000, 1,500 2,000 hours, and so on if necessary. Usually tubes having abnormally long lives are removed from test, as waiting for their failure greatly delays the final results, and the including of a very few abnormally long life figures renders less valuable the average figure obtained.

When only a very few triodes are life tested the average figure obtained may not accurately represent the average life

of a very large number. The Bureau of Standards in its Circular Number 13, covering "United States Government Specifications for Large Tungsten Filament Incandescent Lamps," gives the following tabulation to indicate what they consider an allowable variation from specified life, due to the relatively small number of samples tested:

Number of lamps averaged	Allowable % variation from guaranteed life	Number lamps averaged	Allowable % variation from guaranteed life	Number of lamps averaged	Allowable % variation from guaranteed life
250 and above	5	24–20	12	9	19
249–100	6	19–18	13	8	20
99–55	7	17–16	14	7	21
54–45	8	15–14	15	6	23
44–35	9	13–12	16	5	25
34–30	10	11	17		
29–25	11	10	18		

For each type of triode there is chosen some value for each of the different factors of tube performance which is considered the end of useful operating life. For instance, in the case of Radiotron UV-201-A, the life is usually considered ended when the electron emission has dropped to below 20 milliamperes. Of course, in many circuits the Radiotron is still satisfactory when the emission has dropped to much lower values, but here, again, from the viewpoint of the manufacturer, relative results quickly obtained are of first importance.

In some of the tabulations given in the latter part of the paper it will be noticed that often the life figures for individual tubes are given to an even twenty-five hours. This is owing to the fact that very often between two successive electrical tests some factor such as the emission drops below the prescribed limit and the time at which it passed below this limit must be roughly estimated from the two readings.

Of course, in a certain number of triodes, failure occurs, due to some definite cause, such as filament breakage, air leakage, short circuit of the electrodes, and so on, defects which render them entirely inoperative. In such a case, if the approximate time of this failure has not been noted, a life figure is arbitrarily assigned which represents a time halfway between the last electrical test at which the triode was satisfactory and the test when the defect was discovered.

In general, a complete set of initial readings is taken on all triodes placed on life test so that if during the test some unforseen

peculiarity develops it is possible to make another complete test and see in what way any of the normal characteristics have changed.

Checks are made at least once a day on the various voltages of each section. In order to be sure that there is no defect in the life test equipment so that triodes are not actually operated without plate current, a milliammeter is included in the plate voltage line and its increasing indications are noted as the triodes are one by one placed on life test. As for each type of tube the normal plate current under the life test conditions is known, it is quite easy to determine as the tubes are being placed on life test whether they are operating under approximately the desired conditions.

When an electrical test, in one of the intervals during life, indicates that the electron emission or some other quantity has dropped slightly below the limit assigned for end of life, the triode is returned to life test and continued with the rest of the lot. This is done to obtain information on the further variation of this quantity and also to be certain that the drop was not temporary or due to some inaccuracy in testing. If, at the next test period, the same factor on this triode is still below the allowable limit, the hours up to the previous test are considered the life of the tube.

Results Obtained From Life Testing

(1) *General*

Operation of the life testing equipment for a period of nearly two years has given certain general results which will be first noted:

(a) There are so many factors involved in the life quality of triodes that it is dangerous to extend the results obtained from one test to cover what appears to be a similar case. In other words, there are so many factors, both in manufacture and operation, determining the life of triodes that the same test run six months or a year later very often does not give similar results.

(b) In the case of the development of a new triode involving many new features of design, the causes of failure will go thru a number of different phases before the development is completed. For instance, the very early samples of a new transmitting type may all fail from arcing over of the glass stem near the seal. Triodes in which this defect has been eliminated when put on life test might all fail from filament breakage at a little later period in life, safely going by the period when the first defect occurred.

After this filament breakage difficulty has been eliminated another lot may last slightly longer and then develop difficulty from continued overheating of some part. This is just an imaginary case, illustrating the sort of thing which is quite common in a new development in its early stages.

(c) Seemingly unimportant, or very minute, details of design or manufacture may have a very pronounced effect on life. Sometimes different lots of triodes have surprisingly large life variations and it is often extremely difficult to learn the cause of these differences.

(d) Up to the present time no thoroly satisfactory method for the forced life testing of triodes has been developed which allows the life quality to be judged by a life test covering only a few days. It is believed such a forced life test, that is reliable, can be developed, but a great deal more experimental work will have to be done before it can be used to displace life testing under normal conditions.

(e) To be of value in judging the quality of a product, life testing must be carried on at regular and frequent intervals. Even under these conditions, individual results occasionally will vary widely, due probably to slight variations in the product or method of testing, which have escaped detection, or owing to the fact that the small number of samples chosen for testing happened to be far from representative of the product as a whole.

(f) It is not possible in the present stage of this work to get exact quantitative relations between such factors as life and plate voltage, or some of the varying conditions of use. The best that can be done is to get the general trend of these variations and relative values.

(2) *Specific Results*

Using the life test equipment, which has been briefly described, there have been life tested during the past eighteen months about one thousand lots, comprising nearly 10,000 triodes. Some of these are still on test. In giving some of the actual results obtained the life test summary sheets and work sheets were examined to pick out cases which were typical of clean cut results, indicating certain relations that are of interest. One of the great difficulties of triode life testing is the occasional seemingly contradictory results and the difficulty of analyzing them.

Therefore, in connection with the results given, it must be remembered that these are special cases picked out specifically

to illustrate some point and in most cases are special lots of triodes, representing experimental variations in the standard product.

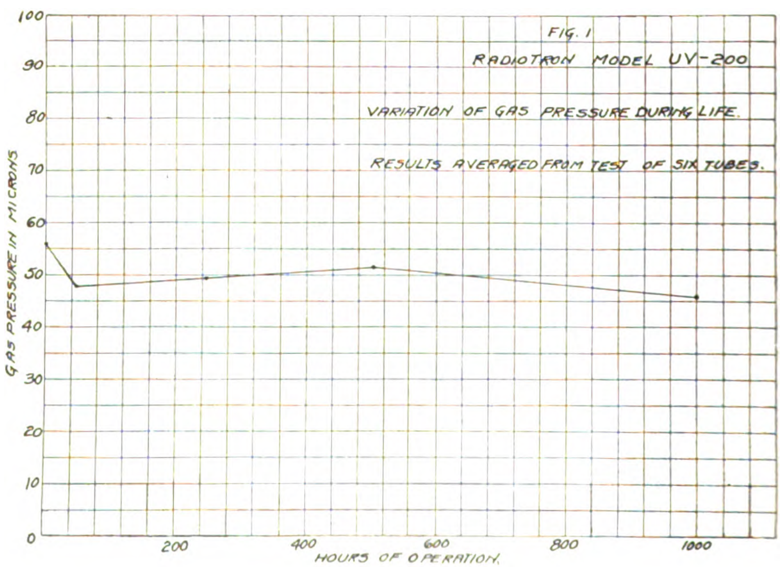
TABLE I

Type of tube tested........................Radiotron UV-200
Lot Number J-M-403
Number of tubes tested................6
Life test conditions:

Filament volts................5.0 A. C.
Plate volts.......................20 D. C.
Grid connected to one end of filament.

Tube Number	Hours Life	Failure due to
4	1300	Filament burn out.
6	1200	Filament burn out.
8	1200	Loss of detector action.
10	1800	Filament burn out.
12	1900	Filament burn out.
15	700	Broken Filament

Average life 1,350 hours. Max. 1,900. Min. 700.



CURVE 1

These illustrate the results of a test run to learn whether the gas pressure in this lot of UV-200 Radiotrons varied during life, and if so in what way.

The gas pressure in these different triodes was initially nearly alike and varied quite similarly during life, so that the average, as plotted, gives a good idea of the result obtained. The curve indicates that the gas pressure during life remains almost constant, with the exception of a rather decided drop near the beginning. The other variations are hardly more than the experimental error in determining the pressure.

TABLE IIA

Type of tube..................Radiotron Model UV-201-A
Lot Number T-S-1
Number of tubes tested................7
Life test conditions:

Filament voltage................5 A. C.
Plate voltage..................60 D. C.
Grid voltage..............5 volts D. C.

Tube Number	Hours life	Cause of failure
*1a	10,000	Base defect
*2a	10,500	Low emission
3a	11,500+	No failure removed from test
*4a	8,300	Low emission
1b	3,800	Low emission
*2b	9,000	
3b	1,000	Accidentally broken

Average life 7,729+. Max. 11,500+. Min. 3,800.

These results were chosen to indicate a rather usual form of variation of electron emission with life on such triodes as the UV-201-A.

As a matter of interest, two different lots were combined; one lot IIA, having exceptionally long life; the other lot IIB, being defective tubes having exceptionally short life. Curve 2 was plotted using the triodes marked with an asterisk on Tables IIA and IIB. Four tubes were picked from each lot that had roughly the same length of life. This was done so that values could be conveniently combined on a percentage basis. The

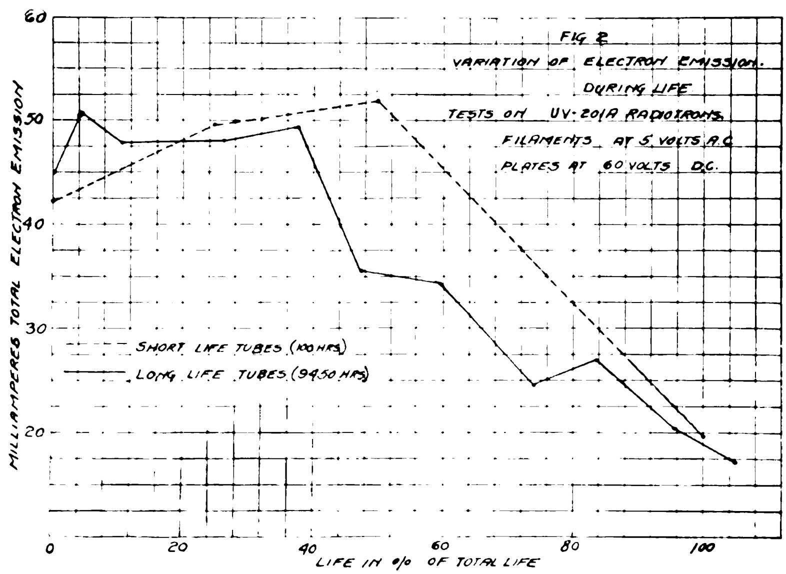
curve shows that there is somewhat of an increase in emission during the first half of life, the latter half of life showing a decrease. In this test the life of the triode was considered as being terminated when the emission dropped below 20 milliamperes.

TABLE IIB

Type of tube.................... Radiotron Model UV-201-A
Lot Number M-W-1 (Defective exhaust)
Number of tubes tested................ 6
Life test conditions:

Filament voltage................. 5 A. C.
Plate voltage.................. 60 D. C.
Grid. . Connected to one filament terminal

Tube Number	Hours life	Cause of failure
*1	75	Low emission
*2	125	Low emission
*3	100	Low emission
*4	100	Low emission
5	75	Low emission
6	75	Low emission



Curve 2

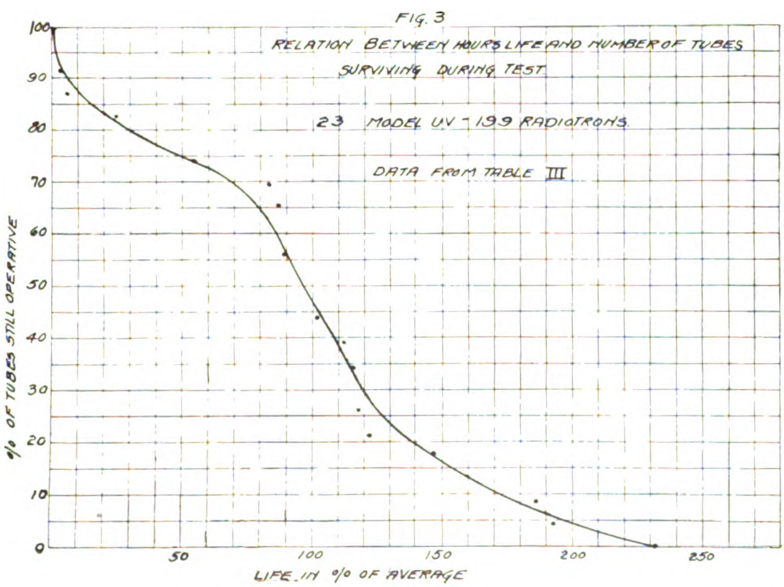
TABLE III

Type of tube....................Radiotron Model UV-199
Lot designation....................N-S-1
Number of tubes tested................23
Life test conditions: (Electrical)

Filament..................3.0 volts A. C.
Plate.....................60 volts D. C.
Grid......Connected to one filament end

Tube Number	Hours life	Cause of failure	Tube Number	Hours life	Cause of failure
4	915	Low emission	32	25	Base defect
8	915	Low emission	33	650	Low emission
21	800	Low emission	34	700	Low emission
22	1,800	Low emission	35	200	Base defect
23	425	Low emission	36	800	Low emission
24	675	Low emission	38	1,250	Low emission
25	1,150	Low emission	39	1,500	Low emission
26	425	Fil burn out	40	700	Low emission
27	25	Short circuit	41	900	Fil. burn out
29	800	Low emission	42	950	Low emission
30	1,450	Low emission	43	875	Low emission
31	35	Broken filament			

Average life 781 hours. Max. 1,800. Min. 25.



CURVE 3

These give the results on a test of twenty-three Radiotrons, Model UV-199. Curve 3 is plotted between life in percentage of average and the percent of triodes still operative. The object of giving the results of this test and plotting this curve is to show that on a relatively large number of tubes the average life is indicated with an approximate degree of accuracy by the number of hours that have elapsed when half of the triodes started on test have failed. The less developed the product the nearer this curve approaches a straight line drawn between the beginning and end of the curve shown. In other words, the more highly developed product shows fewer early failures and fewer exceptionally long lives, the majority failing around average life for the lot.

Table IV is given to illustrate how some very slight detail of manufacture or design will greatly affect the life of the triodes. These four lots were identical, except that the two lots had a very slight difference in the nature of the grid; a difference that would entirely escape ordinary inspection, but happened to be noted in these experimental lots of Radiotrons, Model UV-201-A. All of the life failures on these lots were due to loss of emission, which in connection with the fact that the only difference between the lots was a detail relative to the grids still further emphasises what great care must be employed in Radiotron manufacture and design. All four lots were life tested in identically the same way. In the two lots A and C there were no failures up to 2,000 hours; whereas, in the other two lots, B and D, the average life was low and only one tube lasted a thousand hours.

TABLE IV

Type of tube life tested..........Radiotron Model UV-201-A
Lot Numbers..........R64-A, B, C, D
Number of tubes tested...............36
Life test conditions:

Filament voltage...............5.0 A. C.
Plate voltage.................120 D. C.
Grid voltage..................−6 D. C.

Lots A and C.

All good at the end of 2,000 hours with average emission more than double the lower limit of 20 milliamperes considered the end of life.

Lot B		Lot D	
Tube Number	Hours life	Tube Number	Hours life
1	500	1	175
2	300	2	100
3	300	3	300
4	250	4	275
5	490	5	175
6	350	6	175
7	1050	7	275
8	750	8	200
9	300	All failures low emission. Aver. life 209. Max. 300. Min. 100.	
10	250		

All failures low emission.
Average life 354. Max. 1,050. Min. 250.

In this test, also, end of life was considered as having been reached when the electron emission dropped below 20 milliamperes.

Table V shows the results of a life test conducted on ten Radiotrons, Model UV-204-A, which is a power triode having a rated output of 250 watts.

TABLE V

Type of tube....................Radiotron Model UV-204-A
Number of tubes life tested.............10
Life test conditions: (Electrical)
Plate voltage...........2,000 volts D. C.
Plate current..........0.125 amps. D. C.
Filament..................11 volts A. C.

Object of test—To determine if intermittent filament operation (½ hour on and ½ hour off) was detrimental to life of this particular type of tube.

Lot A—Intermittent operation

Tube Number	Hours life
7,569	1,577
7,572	1,000
7,579	2,500+
7,590	691
7,598	960

Average life 1,346+ hours. Max. 2,500+ hours. Min. 691 hours.

Lot B—Continuous Operation

Tube Number	Hours life
4,137	1,500
7,004	445
7,561	1,370
7,580	291
7,582	1,718

Average life 1,065 hours. Max. 1,718 hours. Min. 291 hours

The object of this test was to determine whether intermittent operation of the filaments greatly shortened the life of this type of triodes. Five of them were operated with the filaments intermittently on one-half hour and off one-half hour. The other five were operated continuously. All other life test conditions were the same. The results indicate that there is certainly in this particular lot of tubes no detrimental effect incidental to intermittent filament operation. The slightly longer life on intermitte nt operation is not noteworthy, because the percentage difference between the lives of the two lots is not greater than would be expected, due to the small number of samples tested.

The commutator described in a previous paragraph was used for automatically turning the filaments on and off during test. In this test the conditions of operations were severe; again, in order to get the answer quickly.

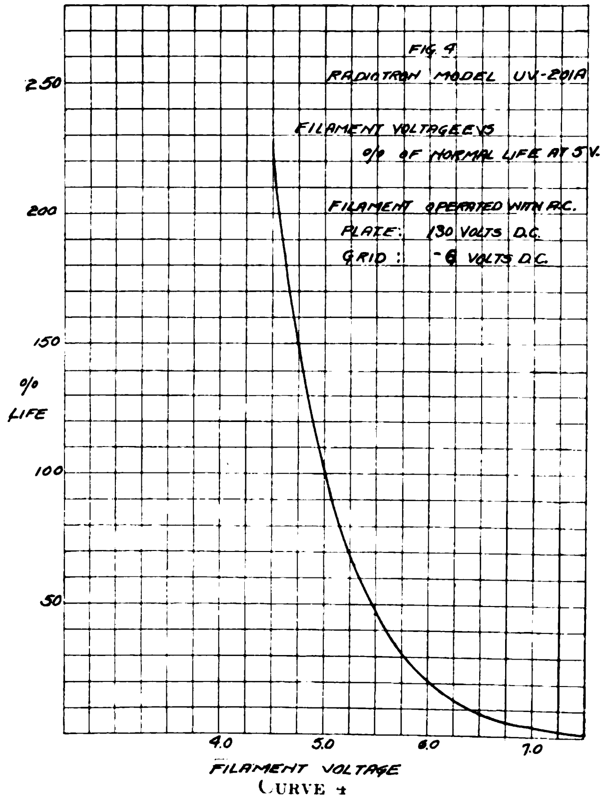
It is believed from some tests made on the filaments of receiving Radiotrons that intermittent operation has no injurious effect. On some of the high power high voltage triodes, however, there is some reason to believe that intermittent operation does have a bad effect on life.

Curve 4 gives the result of a test made on a few Radiotrons, Model UV-201-A, to determine in what way life was affected by variation of filament voltage. The curve indicates that in the range tested the life of this particular lot of tubes was halved or doubled when the filament voltage was increased or decreased ten percent, respectively. Tests have not been run on a sufficient number of tubes to indicate the accuracy of these results, and they are merely given as an example of results obtained from one particular lot.

In making this curve the actual figures were plotted on a separate curve sheet of the semi-log type. A smooth line was

drawn to give the best interpretation of the results and Curve 4 was plotted from this line.

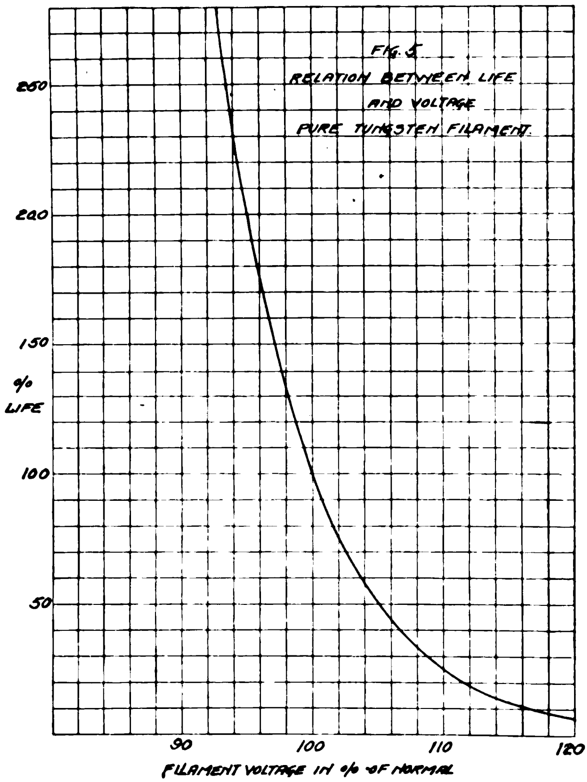
The calculated rate of evaporation of tungsten, confirmed by test experience, indicates that the life of a pure tungsten filament triode, other factors being the same, will be halved or doubled with a five percent increase or decrease, respectively, of filament voltage. This relationship is shown in Curve 5 plotted on a percentage basis.



CURVE 4

In the case of a pure tungsten filament triode the filament temperature is so high that the life is usually terminated by actual burnout and during life there is a slight evaporation of tungsten from the surface so that its diameter is slightly decreased. Experience in the life testing of tungsten filament lamps indicates that on the average the life of a tungsten filament is ended when the evaporation has reached such a point that the diameter has been decreased by about three percent to ten percent, depending

upon the diameter of the filament. The larger the diameter of the filament the greater the percentage of evaporation before failure occurs.



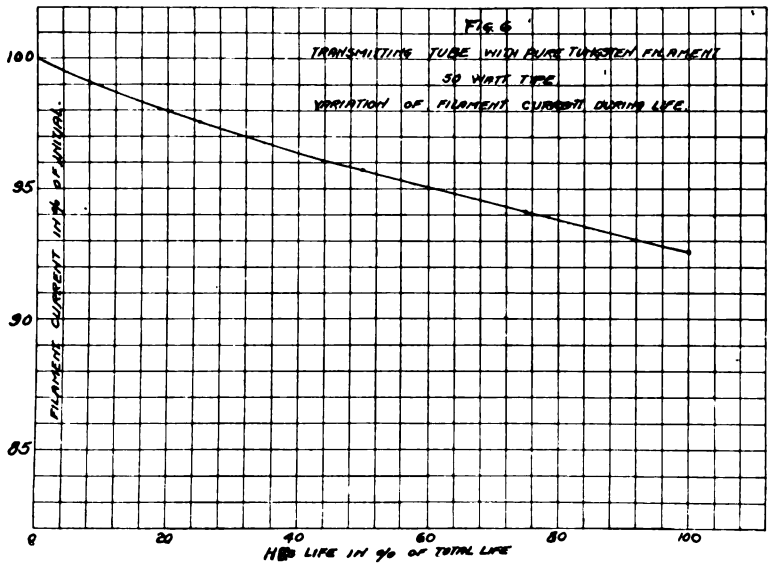
CURVE 5

This decrease in filament diameter during life causes an increase in filament resistance and when the filament is operated at constant voltage there is, therefore, a decrease in filament current.

Curve 6 shows such a filament current decrease, plotted in percent, for a fifty-watt type of triode having a pure tungsten filament.

It might be supposed that this drop of current would cause a drop in electron emission, but this is not always the case, and when it does occur it is often only very slight in amount. The reason for this is that as the current goes down, due to the decrease of filament diameter, so also does the heat radiating sur-

face decrease, and, therefore, the temperature is maintained at nearly constant value. Also, experience has shown that in many cases this evaporation gives a purer tungsten surface which emits electrons at a considerably higher rate per unit of area than the initial surface which might have been somewhat contaminated.



CURVE 6

In conclusion, it is again to be emphasized that these specific results have been selected to illustrate some of the results obtained from the life testing of triodes, but by the very nature of the tests and the way triodes are used, they cannot in any way represent the life to be expected in any individual case of operation.

As explained at the beginning of the paper, the life of a triode is not a constant for any particular type, but is variable as are most of the other characteristics, depending upon many factors.

SUMMARY: It is first pointed out that triodes are life tested primarily as an aid to the manufacturers in proving their performance and useful length of service rather than to obtain any average life figure. The apparatus employed and its method of operation, together with the procedure in handling the data, is next described.

Some actual results obtained are then given to illustrate the methods employed and results obtained. These results are given in the form of tables and curves.

One point emphasized thruout the paper is that triode life is just as much a variable as other factors such as electron emission or impedance.

FURTHER DISCUSSION ON "A METHOD OF MEASURING VERY SHORT RADIO WAVE LENGTHS AND THEIR USE IN FREQUENCY STANDARDIZATION," BY F. W. DUNMORE AND F. H. ENGEL

S. R. Kantebet (by letter): The discussion on the above paper raised by Messrs. Takagishi and Kawazoe has brought out an important point in the distribution of current along a pair of parallel wires carrying stationery waves. The writer himself, in making certain measurements on wave lengths of the order of six meters, has found that wave shapes like the ones experienced by Takagishi and Kawazoe exist even under conditions entirely different from theirs.

In these experiments a 50-watt Mullard valve was used as the source of short radio waves. The oscillatory circuit was of the Hartley type. A single turn of number 14 standard wire gauge 3 cm. in diameter formed the plate circuit coil. In the grid circuit there was a coil of 3 turns of 1 cm. diameter, mounted at a distance of about 6 cm. from the plate coil. The parallel wires were bent at the input and into a loop of about 20 cm. diameter. The loop was nowhere nearer the oscillatory circuit than about 30 cm. Finally, instead of a heavily shunted thermal instrument, a selected carborundam crystal connected onto a galvanometer was used for measuring the current distribution.

It will be seen from this that the coupling between the wires and the valve circuit could not be called tight, especially having regard to the fact that the oscillatory circuit was so small, the parallel wires so far off, and the the input to the valve never more than about 25 watts. Still when current measurements were made with the crystal and galvanometer for a distance covering over three half-waves along the wires, and the results plotted to a distance base, it was found that each half-wave had two maxima with a central depression. And this distribution of current persisted even when the coupling loop was distorted into a thin rectangle. Hence the objection of tight coupling, as suggested by Dunmore and Engel, does not seem to answer the point.

It will be interesting to note here that whereas this distribution was got by chance, Messrs. G. Lamm and E. Graham have

obtained similar curves in an entirely different way ("Wireless World and Radio Review,"—December 31, 1924 and January 7, 1925.) Their parallel wires were 16 meters long, 2 mm. in diameter with a clearance of 5 cm. Measurements were made with a thermo-ammeter of 60 ohms resistance. However, in contrast with other experiments, they short-circuited the wires at one end and induced the radio frequency currents nearer the other end of the wires which were here joined together thru an electrolytic resistance the value of which (117.5 ohms) was got at experimentally so as to give a very pronounced central depression for every half wave. They give 500 ohms as the surge impedance of these wires.

In explanation of these curves, they say: "It will, of course, be understood that the influence of the measuring instrument (on the shape of the wave) will increase as its impedance becomes smaller compared with the characteristics of the parallel wires, their characteristics being 500 ohms for these frequencies (30 thousand kilocycles). We can connect a resistance between the terminals at end B and carry out similar calculations. The result will be that we can obtain any form of curve from a pure sine wave.

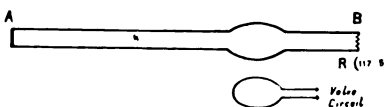


Fig 1

"If we want to get the original curve, the impedance of the measuring instruments should be great compared with the characteristic of the wires." They also prefer measurements being made with nodes instead of loops for reference points, as they say high accuracy is possible there.

The central depression in the INSTITUTE experiments was attributed to the presence of the 3rd harmonic component in the fundamental wave, without experimental verification. So much for the central dip in the distribution curves.

A word may here be said about the different means available for measurement, some of these being a thermal instrument, with or without a shunt, a crystal rectifier and a valve detector. The latter two seem to be specially suitable from the point of view of reflection. Pronounced current loops exist on a pair of wires only when the co-efficient of reflection,

$$K = \frac{R - Z_0}{R + Z_0}$$

is unity. The resistance of crystals and thermionic valves is several thousand ohms, whereas the surge impedance of parallel wires is usually a few hundred ohms. Consequently the reflection factor becomes practically unity (+1) and not very different from that of the method under review which works with $K = -1$.

Coming to the thermal instruments, the following doubt arises. Firstly, they are not very sensitive for small currents and their scales are crowded near the zero point. So one has only to work at the current loops. In the INSTITUTE experiments, as also those referred to above, it has been found preferable to work at the node. In the former case, the thermal instrument gave a much flatter curve with not very definite maxima and minima. The crystal, on the other hand, showed capacity for considerable precision, the node having been traced down to a few micro-amperes.

Secondly, the use of a shunt across the thermo-ammeter seems to be open to the following objection. As has been admitted ("Scientific Papers of the Bureau of Standards," number 491), almost all the current passes thru the shunt and only about a four-hundredth part goes to the heater of the instrument. Suppose that, for example, for a certain position 100 microamperes flow thru the heater, then 40,000 flow through the shunt. Suppose next that, due to a certain shifting, 1 percent change takes place in the current. Out of a total change of 401 microamperes, the current change in the heater will be only 1 microampere in the hundred already flowing thru it. Consequently the change in the heating effect is in the proportion of 10^{-4}, which should be very difficult indeed to perceive. With the shunt removed, all the change would affect the heater and, as such, would be easily perceivable. So the heavy shunt seems to reduce the sensitivity of the instrument very considerably.

As regards reducing the effective resistance of the wire system thru the use of the shunt, perhaps the object could be achieved by using a thicker wire without losing the sensitivity of the instrument.

Department of Electrical Technology,
Indian Institute of Science,
Bangalore, India.

F. W. Dunmore, F. H Engel, and **A. Hund** (by letter): In reply to Mr. S. R. Kantebet's remarks, the parallel wire method of measuring very short wave lengths described by F. W. Dunmore and F. H. Engel in the PROCEEDINGS OF THE INSTITUTE OF

Radio Engineers, volume 11, number 5, article 23, page 407, and discussed theoretically by Dr. A. Hund in "Bureau of Standards Scientific Paper" number 491, "Theory of Determination of Ultra-radio Frequencies by Standing Waves on Wires," has since been further investigated experimentally by other members of the Bureau of Standards radio laboratory staff. With the conditions as described in detail in the above papers, it was never possible to obtain two maxima as obtained by Mr. K:antebet. The resonance settings were always decidedly sharp. The resonance settings were not nearly so sharp when no shunt was used on the indicating instruments. With the shunt, check settings could be made within 1 mm., whereas without the shunt it was impossible to duplicate settings within 2 mm. Resonance curves taken with and without the shunt showed much sharper resonance with the shunt. The experiments here agree with Mr. Kantebet's suggestion that a crystal detector would be a good means of indication. A crystal detector in series with a portable galvanometer (full-scale deflection = 1 milliampere) as the indicating device was found quite satisfactory. A shunt was used across this combination. The use of this combination increased the sensitivity. As brought out in the above paper, the indicating instrument bridging across the parallel wire system should be as compact as possible. It is doubtful whether reliable results can be obtained when long leads run to a tube or any other indicator. The parallel wire system should be suspended freely as far away as possible from other apparatus.

Radio Laboratory of Standards,
Department of Commerce,
Washington, D. C.

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

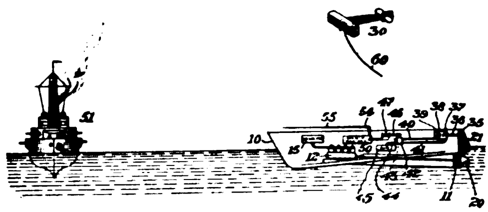
ISSUED JULY 7, 1925—AUGUST 25, 1925

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,544,746—J. H. Hammond, Jr. Original filed August 10, 1916, issued July 7, 1925.



NUMBER 1,544,746—Method and Apparatus for Controlling Water Craft from Aircraft

METHOD AND APPARATUS FOR CONTROLLING WATER CRAFT FROM AIRCFAFT, where the water craft is provided with steering means and a radio receiving apparatus which is arranged to control the steering means. A directional loop is arranged aboard the water craft so that signals may be received from a radio transmitter aboard an aircraft directly over the loop for controlling the steering mechanism aboard the water craft.

1,545,040—W. Dornig, filed May 10, 1923, issued July 7, 1925.

MULTIPLYING TRANSFORMER for radio transmission systems, in which a secondary oscillation circuit is connected between the transformer and the antenna system, and the circuit tuned to the frequency of the antenna system. An auxiliary circuit is connected to the transformer and tuned to the upper harmonics of the primary frequency by which the desired frequency is impressed upon the antenna system.

1,545,041—W. Dornig, filed May 10, 1923, issued July 7, 1925.

CIRCUIT ARRANGEMENT FOR FREQUENCY-MULTIPLYING

*Received by the Editor, September 15, 1925.

TRANSFORMERS, wherein high frequency oscillations are produced in a single transforming step by wave distortion in the transformer core. A circuit is provided including with oscillation producer which is arranged to impress upon a radiating system the desired high frequency signaling energy due to the exclusion of undesired frequencies.

1,545,207—C. G. Smith, filed August 30, 1920, issued July 7, 1925. Assigned to S-Tube Corporation, Dover, Delaware.

ELECTRICAL APPARATUS comprising a tube construction having an electrode of extended surface area with a second minimum area across an intervening space. The space between juxtaposed surfaces of the electrodes is sufficiently short and the gas pressure sufficiently low as to prevent initiation of substantial conduction directly across the space by potentials high enough to initiate conduction across longer gaps. The electrodes are insulated to prevent substantial conduction between all other portions of the area thereof. The plates form a condenser disposed in vacuum with extremely high break-down potential.

1,545,247—J. O. Gargan, filed October 17, 1919, issued July 7, 1925. Assigned to Western Electric Company, Incorporated.

SUPPORT FOR VACUUM TUBES, by which vibrations are prevented from reaching the tubes by means of a resilient mounting. The vacuum tube support comprises a layer of yielding material, such as sponge rubber, with a rigid member attached to be spaced from the material with the vacuum tube secured to this member. The tube projects from the member and toward the material. A pluurality of tubes are suspended from the flexible mounting by members which hang downward from the resilient material.

Re. 16,113—R. J. Fitzgerald. Original filed November 4, 1919, issued July 14, 1925. Assigned one-half to J. Arthur Fischer, New York.

CONDENSER having a plurality of stationary conducting plates and a plurality of interposed movable plates connected together and movable in their own planes into and out of position between the stationary plates. Each plate has a non-conducting coating on opposite sides thereof to enclose the plate. The coating of each movable plate is separate from the coating of each adjacent stationary plate.

1,545,502—Marius Latour, filed August 29, 1921, issued July 14, 1925. Assigned to Latour Corporation.

RADIO TELEPHONY, in which telegraphic signals are received from a radio frequency transmitting station by super-imposing locally generated radio frequency oscillations on the receiving oscillation and adjusting the difference of the locally generated frequency and the signal frequency within the limits of audibility so that the beat note does not interfere with the speech. The invention is an extension of the heterodyne principle used in telegraphic systems to reception in radio telephony.

1,545,523—H. Riegger, filed November 18, 1924, issued July 14, 1925. Assigned to Siemens and Halske, Berlin.

MEANS FOR TRANSMITTING TIME SIGNALS, by the discharge of a condenser across an inductance and an ohmic resistance both of which are so dimensioned that the discharge requires a certain duration only which is determined by the prescribed accuracy. The condenser circuit is accurately adjusted so that the discharge occurs over predetermined time intervals.

1,545,591—J. J. Madine, filed March 6, 1923, issued July 14, 1925. Assigned to Western Electric Company, Incorporated.

MANUFACTURE OF ELECTRON DISCHARGE DEVICES, in which a tight seal between the leading-in wires and the glass envelope is insured. The envelope of the tube and the glass stem which supports the electrode are made of different types of glass. Apertures are formed at the junction of the different types of glass and the leading-in wires passed therethru by applying a bead of glass to the leading-in wires. The bead is fused to close the aperture between the glass portions.

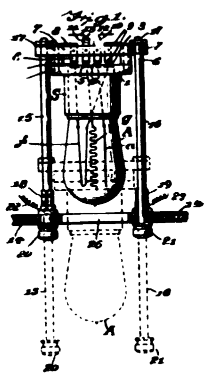
1,545,599—P. O. Pedersen, filed March 22, 1921, issued July 14, 1925. Assigned to Poulsen Wireless Corporation.

METHOD OF AND MEANS FOR PRODUCING OSCILLATING CURRENTS OF RADIO FREQUENCY, in which the efficiency of an arc generator is increased. The arc generator is provided with two electrodes between which the arc is formed with magnetic pole pieces arranged on opposite sides of the arc with adjustable shoes on the pole faces for shaping the magnetic field about the arc. The arc voltage is maintained at a low value during the longer part of the cycle.

1,545,607—J. C. Schelleng, filed December 22, 1920, issued July 14, 1925. Assigned to Western Electric Company, Incorporated.

System of Wave Distribution, in which circuits are provided for maintaining the frequency of an oscillator constant under varying conditions of load. A wave distorting circuit is coupled to a frequency determining circuit and connected to a source for producing harmonics of the wave from the source. A work circuit is provided which constitutes parallel resonant paths for the harmonic frequency which is maintained constant and which may be impressed upon an antenna circuit for transmission.

1,545,639—S. Cohen, filed July 11, 1922, issued July 14, 1925. Assigned to Grace A. Barron, New York.



Number 1,545,639 —Vacuum Tube Mounting

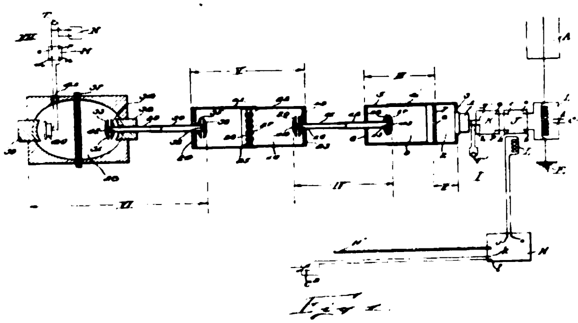
Vacuum Tube Mounting, in which the electron tube is carried upon a movable structure by which the tube may be normally disposed within a cabinet and withdrawn to permit removal or re-insertion of a tube in the socket member. The socket member is guided by rods on opposite sides thereof enabling it to be readily moved to a point adjacent the front of the panel or returned to a position behind the panel when the tube is to be inserted or removed from the socket.

1,545,654—A. H. Hoppock, filed November 20, 1920, issued July 14, 1925. Assigned to Western Electric Company, Incorporated.

Water-Cooled Anode for Vacuum Tubes, comprising a closed vessel having thin parallel side walls closely adjacent each other. One of the side walls is provided with depressions form-

ing a tortuous path thru the vessel in which a cooling fluid may be passed.

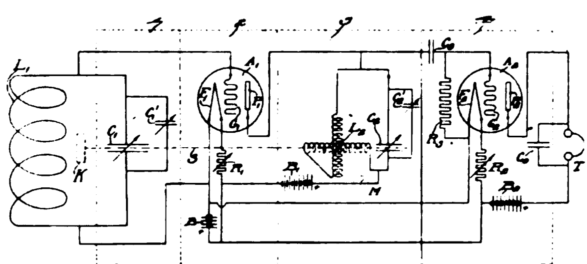
1,545,697—O. C. Roos, filed November 4, 1921, issued July 14, 1925.



NUMBER 1,545,697—Electromagnetic Wave Receiving System

ELECTROMAGNETIC WAVE RECEIVING SYSTEM, in which the effect on the signal-indicating device of electrical vibrations created in the system by static disturbances may be acoustically eliminated while the desired signals may be amplified and received. A percussion chamber is provided into which the sound vibrations from the telephone receiver of the receiving system are delivered. A reverberation chamber is acoustically associated with the percussion chamber with which there is associated a stationary wave separating chamber from which desired signaling energy is delivered to a signaling indicating device while the passage of static disturbances is acoustically prevented.

1,545,940—C. Cabot, filed May 25, 1922, issued July 14, 1925.



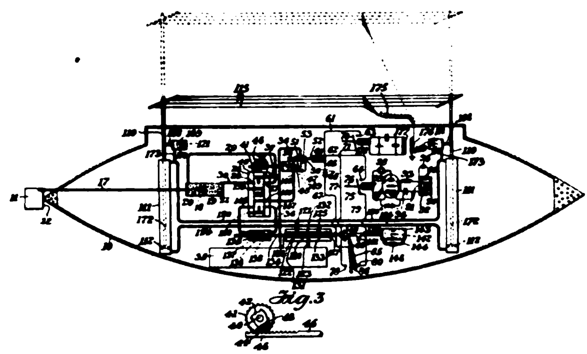
NUMBER 1,545,940—Electromagnetic Wave Receiving System

ELECTROMAGNETIC WAVE RECEIVING SYSTEM of the radio frequency amplifier type which is non-regenerative. The radio

frequency amplifier comprises several stages having parallel branch circuits with inductance in one branch and a capacity in another branch, both of which may be simultaneously adjusted to resonance. A vernier tuning adjustment is provided for insuring the exact resonance of the system and preventing oscillations in the radio frequency amplifier.

1,546,579—J. H. Hammond, Jr., filed October 14, 1925, issued July 21, 1925.

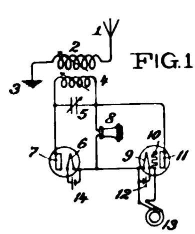
Fig. 1



Number 1,546,579—Dual System of Control for Dirigible Devices

Dual System of Control for Dirigible Devices, such as marine vessel, a torpedo, an airplane, or any other dirigible device, which may be steered or stabilized with respect to a given axis. A control mechanism is provided on the dirigible device whereby the receiving antenna may be automatically elevated on receipt of a particular signal.

1,546,639—C. L. Farrand, filed May 14, 1919, issued July 21, 1925. Assigned one-third to Cornelius D. Ehret, Philadelphia.



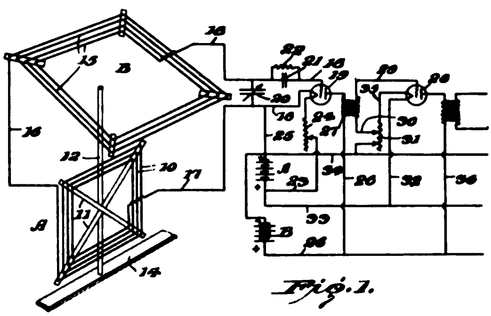
Number 1,546,639—Method of and Apparatus for the Reception of Radio Signals

METHOD OF AND APPARATUS FOR THE RECEPTION OF RADIO SIGNALS, by impressing the received signaling energy upon the anode-cathode circuit of a thermionic impedance. The impedance is independently varied by varying an electrostatic field at predetermined frequency different from the incoming signaling frequency whereby the signaling current may be translated into intelligible sounds.

1,546,696—J. F. Yates, filed August 7, 1923, issued July 21, 1925.

VACUUM TUBE, in which a mid-tap is taken from the cathode to a contact exterior of the tube which connects to the grid circuit and alternating current supplied across the cathode. The variations in the input circuit therefore occur between a substantially equipotential point on the cathode and the grid, avoiding interference from the alternating current hum.

1,546,731—J. H. Herzog, filed January 23, 1923, issued July 21, 1925. Assigned to Herzog Radio Corporation.



NUMBER 1,546,731—Radio Apparatus

RADIO APPARATUS, in which signals are received from a single direction only and undesired signals from the opposite direction blocked out. A pair of coil antennas is arranged in such physical relationship and electrically connected in series in such manner that uni-directional reception is secured.

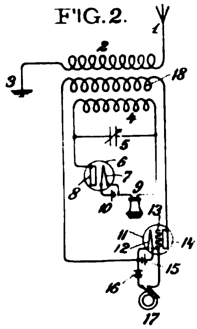
1,546,776—W. R. Bullimore, filed November 24, 1924, issued July 21, 1925.

MANUFACTURE OF FILAMENTS FOR ELECTRIC LAMPS, THERMIONIC TUBES, AND THE LIKE, which consists in passing a base or metallic core thru a liquid containing in suspension an agglu-

tinant and an alkaline earth metal compound for the purpose of simultaneously effecting a coating of agglutinant and metal compound. The agglutinant is subsequently burnt away, leaving the filamentary cathode.

1,546,781—C. L. Farrand, filed October 13, 1919, issued July 21, 1925. Assigned one-third to Cornelius D. Ehret, Philadelphia, Pennsylvania.

FIG. 2.



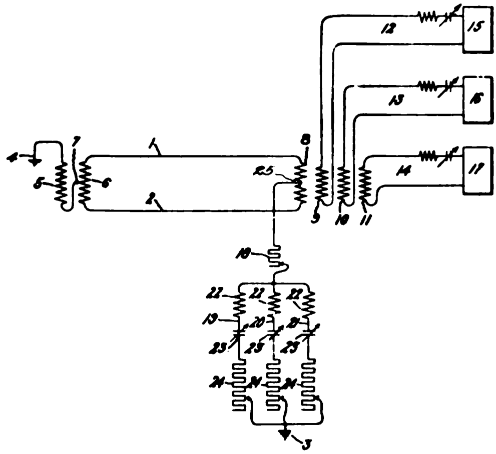
NUMBER 1,546,781 —Methods of and Apparatus for the Reception of Radio Signals

METHOD OF AND APPARATUS FOR THE RECEPTION OF RADIO SIGNALS transmitted by continuous waves where the received signaling energy is impressed upon a thermionic detector, the operation of which is periodically varied by means of a thermionic impedance which varies in magnitude periodically at a frequency above audibility. The received signaling current as modified by the impedance variation is translated into intelligible signals.

1,546,801—C. P. Sorensen and R. W. Satterholm, filed January 22, 1915, issued July 21, 1925.

HOLDER FOR CONDENSERS, which does not require the use of soldered connections between the condenser terminals and the metallic clamps on the condenser stack. A pair of resilient arms extend on opposite sides of the clips which clamp the condenser stack and support the condenser mechanically at the same time that good electrical connection is established.

1,546,878—E. F. W. Alexanderson, filed June 7, 1921, issued July 21, 1925. Assigned to General Electric Company.


Number 1,546,878—Radio Receiving System

Radio Receiving System for the operation of a multiple number of receivers simultaneously on the same antenna system. A long horizontal receiving antenna is used with a plurality of receiving sets connected therewith and a plurality of tuned circuits connected in the ground circuit for reflecting over the antenna from another point therein currents of equal magnitude and opposite phase to the undesired currents for suppressing undesired oscillations in the system.

1,547,412—Roy Crocker, filed May 8, 1922, issued July 28, 1925

Variable Condenser, in which flexible metallic sheets spaced by flexible dielectric material are rolled from one roller to another for effectively varying the capacity of the condenser.

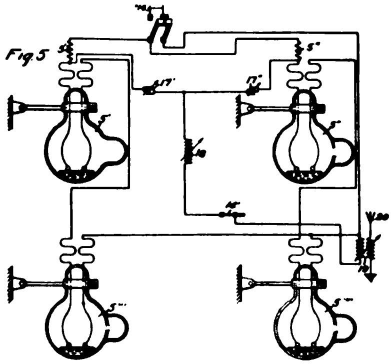
1,547,670—J. W. Radu, filed November 8, 1910, issued July 28, 1925. Assigned to Western Electric Company, Incorporated.

Vacuum Tube, in which the base of the tube is formed from a molded piece of insulated material with dome shaped projections on the base thereof over which metallic contacts are inserted and connected to the electrodes of the tube thru the molded base. This is the construction for the terminal contacts on the Western Electric Company's "peanut tube."

1,547,684—H. C. Rentschler, filed August 17, 1923, issued July 28, 1925. Assigned to Westinghouse Lamp Company.

Oscillation Generator and Joint Operation Thereof, where oscillation generators of the arc type are employed in asso-

ciation with a transmitting antenna system. The arcs are enclosed in vessels which contain a pool of mercury. The vessels are tilted in order to start the arcs, thereby bridging the distance between the arc electrodes with the mercury, and while in this position the power is impressed across the electrodes and then the vessels tilted in upright position for setting up a sustained arc.



NUMBER 1,547,684—Oscillation Generator and Joint Operation Thereof

1,547,753—W. G. Housekeeper, filed August 30, 1920, issued July 28, 1925. Assigned to Western Electric Company.

PROCESS OF TREATING METAL to a high heat in a vacuum by which the metal is prepared for use in thermionic tubes by clearing the metal of occluded gases. The electrodes for the tubes are stamped from untreated metal and the formed electrodes then subjected to a high heat in vacuum. The electrodes are assembled within the enclosing vessel and then the vessel evacuated simultaneously with the heating of the electrodes.

1,547,760—R. W. King, filed July 22, 1919, issued July 28, 1925. Assighed to Western Electric Company, Incorporated.

ELECTRON DISCHARGE DEVICE, in which provision is made for preventing leakage between the leading-in wires at high voltages and ionization of the gas in the tube at high voltages due to the emission of occluded gases from the electrodes is avoided. The electrodes are supported within the tube on arms of refractory material. Wires are sealed into the tube and a metallic clamp provided between the ends of the arms and the wires.

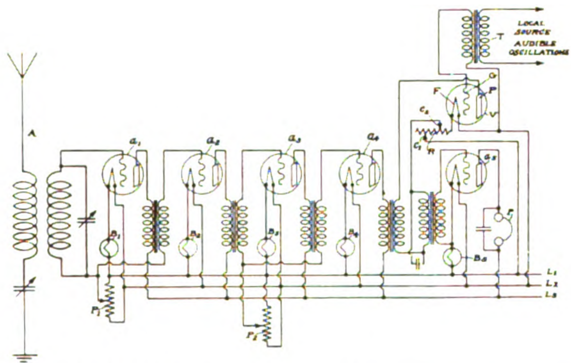
1,547,812—W. F. Hendry, filed September 4, 1919, issued July 28, 1925. Assigned to Western Electric Company, Incorporated.

Vacuum Tube and Method of Manufacturing the Same, in which the electrodes within the tube are positively spaced apart by an insulating block of fusible material which is poured into a mold into which supporting wires from the electrodes extend. When the material hardens, the electrodes are positively secured against displacement.

1,547,885—J. F. Lindberg, filed December 19, 1923, issued July 28, 1925. Assigned to Reliance Die and Stamping Company.

Electric Condenser of variable construction in which the rotatable plates are interleaved with the stator plates in varying degrees with a coarse adjustment control shaft and a fine adjustment cam and lever mechanism actuated by a second concentric shaft by which the rotor plates may be moved thru small angular increments for finely adjusting the capacity of the condenser.

1,547,995—W. L. Carlson, filed February 14, 1922, issued July 28, 1925.



Number 1,547,995—Method of and Apparatus for Receiving Radio Signals

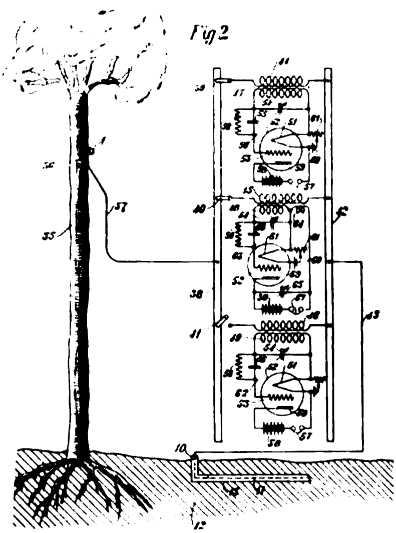
Method of and Apparatus for Receiving Radio Signals, which consists in rectifying the radio frequency signaling currents in a circuit in which no current can normally flow and producing variations of impedance in the circuit only when the signal frequecy is present and then translating the rectified current as modified by the impedance variations into intelligible signals. One of the tubes in the system has its plate filament circuit interposed in the coupling circuit with the preceding tube while a local source of audio freqency oscillations is interposed in the grid filament circuit and a negative bias provided on the plate

electrode of the tube. The tone of the received telegraphic signals may be readily adjusted at the rceiver.

1,548,015—A. B. Bergen, filed October 4, 1924, issued July 28, 1925. Assigned to United States Tool Company, Incorporated.

Condenser Plate System, wherein the stator plates of a variable condenser are made up of a continuous strip of sheet metal folded back and forth from one end to the other. Narrow strip-like tongues of sheet metal integrally connect adjacent edges of the stator plates, providing positive spacing means for the plates and electrical connectors between which securing posts may be inserted. This method of manufacturing a variable condenser facilitates the quantity production of condensers.

1,548,032—G. O. Squier, filed August 8, 1919, issued August 11, 1925.



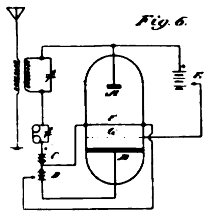
Number 1,548,032—Tree Telephony and Telegraphy

Tree Telephony and Telegraphy, where living vegetable organisms are employed as a radiating system for a radio transmitter. The output of a radio transmitting set is connected to a point in a tree or other vegetable organism which serves as an antenna.

1,548,408—N. B. Davis, filed October 18, 1924, issued August 4, 1925.

CRYSTAL DETECTOR, in which a body of mercury is provided with a crystal floating thereon and a vertically adjustable contact member arranged to contact with the top surface of the crystal. The detector is arranged for vertical mounting on the front of a panel of a radio receiver.

1,548,757—J. Scott Taggart, filed November 30, 1920, issued August 4, 1925. Assigned to Commercial Cable Company.



NUMBER 1,548,757—
Electron Discharge
Device

ELECTRON DISCHARGE DEVICE, which includes a cathode and two anodes. A cathode is provided and the circuits arranged so that electrons flow to both of the anodes. The circuit is such that the potential of one of the anodes may be varied with respect to the cathode and the flow of electrons to the anode may be increasingly diverted to the other anodes as the potential of the first named anode is increased. The tube may be used as an amplifier and is termed in the specification "a negatron."

1,548,801—O. B. Jacobs, filed March 24, 1922, issued August 4, 1925.

VARIABLE CONDENSER construction comprising inter-leaved flexible condenser plates with interposed flexible dieletric interfolded along parallel lines and longitudinally adjustable with respect to each other. The plates may be tubular in form and may slide with respect to each other or a flat or rectangular shape of plate may be used, the plates being telescopically related for varying the capacity of the condenser.

1,548,811—J. H. Hammond, Jr., filed January 6, 1920, issued August 4, 1925.

SYSTEM OF CONTROL BY LIGHT WAVES, wherein a circuit arrangement is provided for preventing the continued conductivity

of a solenium cell circuit after the cessation of the light signal which normally results in a considerable lag in the circuit. The received light energy is concentrated upon a solenium cell and then the circuit to the cell momentarily opened as a result of the received signal for reducing the inertia of the operation of the cell as a control element in the receiving system.

1,549,183—G. H. Clark, filed February 23, 1921, issued August 11, 1925. Assigned to Radio Corporation of America.

RADIO SIGNALING APPARATUS, consisting of an arc generator which does not radiate between the signaling periods. An electromagnetic mechanism is provided on the arc so that the arc electrodes, which are normally short-circuited, are separated for the production of an arc during the signaling period by which oscillations are set up and impressed upon the antenna system. The arc is intermittently suppressed and re-ignited for the production of signals.

1, 549, 253—W. G. Housekeeper, filed October 5, 1920, issued August 11, 1925. Assigned to Western Electric Company.

ELECTRODE FOR ELECTRON DISCHARGE DEVICES, which consists of a boxlike grid structure which entirely surrounds the cathode adjacent the plate electrode. The grid structure is stamped from sheet metal with upwardly extending side members and lateral members integral with the side members. The lateral members are bent in opposite directions to form a rectangular boxlike structure around the cathode.

1,549,310—Leroy E. Humphries, filed October 22, 1923, issued August 11, 1925. Assigned one-fourth to Asa W. Candler, Atlanta, Georgia.

RADIO FREQUENCY SIGNAL RECEIVING SYSTEM for reduction of strays and interference from undesired signals. A pair of electron tube systems is provided which connects to separate collecting circuits tuned to different radio frequencies. The output circuits of the electron tube systems are tuned to substantially the same frequency. A link circuit is provided for coupling the two electron tube systems upon which link circuit local oscillations are impressed and delivered to each of the electron tube systems. The frequency from the local oscillator, which is impressed upon one electron tube system, is equal to the frequency of the opposite radio frequency energy collecting circuit, and vice versa. The output circuits are differentially coupled and arranged

to deliver a resultant field to an independent radio receiving circuit.

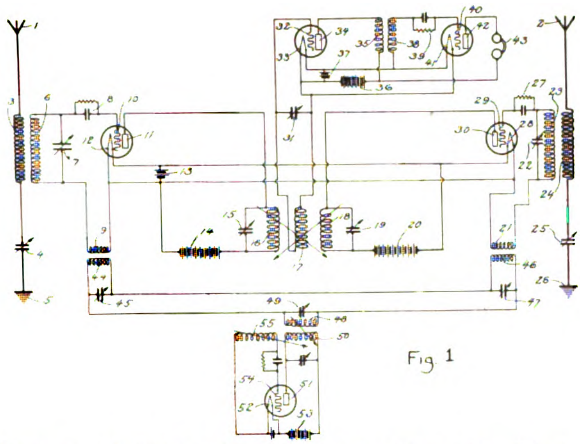
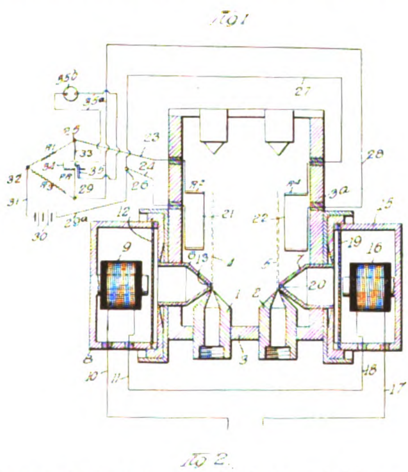


Fig. 1

NUMBER 1,549,310—Radio Frequency Signal Receiving System

1,549,196—R. E. Hall, filed November 14, 1918, issued August 11, 1925. Assigned to Hall Radio Corporation.



NUMBER 1,549,196—Method of and Means for Translating Sounds

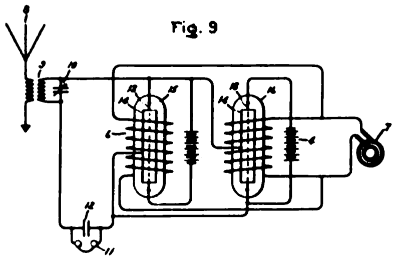
METHOD OF AND MEANS FOR TRANSLATING SOUNDS, comprising jet relays arranged in the output circuit of a radio receiving circuit and selectively responsive to a particular tone. The jets operate alternately to heat and cool a resistance unit which is connected to control a suitable recording circuit. The jets are

arranged to be extremely sensitive to a particular tone quality of the desired signal.

1,549,355—J. J. Chegan, filed February 12, 1924, issued August 11, 1925.

Vacuum Tube Contact, which consists of a coil spring secured to an inwardly extending strip-like member and to which the pins on the base of an electron tube may be inserted. The coil springs insure good contact with the pins of the electron tube over a large area of the pin.

1,549,737—E. F. W. Alexanderson, filed September 15, 1919, issued August 18, 1925. Assigned to General Electric Company.



Number 1,549,737—Signaling System

Signaling System for radio reception where a detector is provided which consists of a resistance device which may be included in the receiving circuit. The current thru this device substantially follows Ohm's law and is, therefore, directly proportional to the impressed voltage. The necessary asymmetry of the device for securing the required rectifying effect is secured by periodically varying the value of this resistance by means of a suitable modifying force which is controllable at the will of the operator at the receiving station. The value of the resistance is controlled or varied in such a way that it is made comparatively small during desired periods and is made exceedingly large during other periods. This permits the flow of an appreciable current in the receiving circuit during desired portions of the impressed signaling wave and causes the practical suppression of the current in receiving circuit during other portions of the signaling wave.

1,549,882—L. L. Jones, filed June 7, 1924, issued August 18, 1925. Assigned to Amsco Products, Incorporated.

Condenser of the variable type in which end supporting plates are provided for enclosing a set of stator plates. A pair of transversely arranged insulating beams are carried by each of the end plates in a plane normal to the plane of the end plates and which serve to support the stator plates. The rotor plates are journaled in the end plates in such manner that the rotor plates may be interleaved between the stator plates.

1,549,926—F. Schneider, filed November 16, 1921, issued August 18, 1925.

Receiving Device for Electric Waves, consisting of a crystal detector where the crystal member is secured between two plate electrodes which may be adjusted with respect to the crystal for securing a junction of maximum sensitivity.

1,550,016—W. W. Dodge, Jr., filed October 19, 1922, issued August 18, 1925.

Electrical Condenser, in which coarse and fine adjustment of variable condenser plates is secured by a mechanism under control of a single knob or actuator. The actuator is positively connected to a movable condenser plate or plates of small capacity with a lost motion connection between the small unit and the large unit so that the actuator may be turned in one direction to engage the large unit, and by continued movement in this direction will then adjust both units in unison. Reverse movement of the actuator within the limits of the lost motion will then produce a relatively negative adjustment of the small unit only.

1,550,421—L. A. Bonish, filed April 27, 1922, issued August 18, 1925.

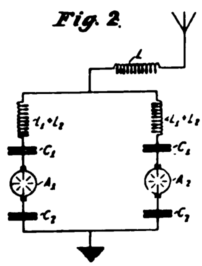
Detector of the crystal variety in which a crystal is supported on the end of a resilient spring arm which is arranged to yield vertically at its free end. A cat whisker engages the crystal and means are provided for adjusting the position of the crystal and the position of the cat whisker with respect thereto.

1,550,571—H. J. Round, filed March 31, 1920, issued August 18, 1925. Assigned to Radio Corporation of America.

Receiving System for Radio Telegraphy and Telephony, in which the effect of certain signals upon a receiver may be eliminated by hetrodyning the signals in each of two receiving systems which are spaced a fraction of the wave length of the signals apart. The same phase difference between the signal

and the heterodyne may be established and therefore the beat currents in the two antenna systems are made to balance one another with the result that the signals are ineffective. The heterodyne energy may be supplied from a separate antenna system set up with relation to the receiving antenna so that the desired phase relation is brought about for the reception of one of the signals to the exclusion of another.

1, 550, 682—H. M. Dowsett, filed September 23, 1922, issued August 25, 1925. Assigned to Radio Corporation of America.



NUMBER 1,550,682—
Arc Generator of
Electric Oscillations

ARC GENERATOR OF ELECTRIC OSCILLATIONS where two arcs are operated in parallel with minimum inter-arc current circulating between the arcs for affording the maximum output current in the antenna system. The inter-arc circuit is provided with condensers of large capacity as compared with the output circuit. The output of each arc is fed into the antenna system with substantially no circulating current in the inter-arc circuit.

1,550,768—H. W. Weinhart, filed July 14, 1919, issued August 25, 1925. Assigned to Western Electric Company.

ELECTRIC DISCHARGE DEVICE, in which the electrodes are supported on wire members which extend from the glass press within the tube. A cylindrical anode is provided with spacer members thereon which engage with the inner walls of the tube for preventing movement of the electrodes. The grid electrode has each turn thereof fastened to the supporting rod which extends from a glass press.

1,550,877—E. L. Chaffee, filed March 10, 1916, renewed September 11, 1923, issued August 25, 1925. Assigned to John Hays Hammond, Jr.

Electric Relay operated from a detector with a circuit arrangement for increasing the sensitiveness of the detector so that it is normally adjusted in such manner that no current flows in the relay circuit. The signaling energy is employed for controlling the current in the output circuit of the tube, the curcuit being so proportioned that the persistence of the operation of the detector, after the impulse which initiated such operation has ceased, is reduced to a minimum.

1,551,087—F. Christiani, filed May 17, 1924, issued August 25, 1925.

Radio Receiving Set, having a crystal detector therein mounted upon a panel on one side of which a variable inductance is variously mounted, and on the other side of which a cover portion is provided for entirely enclosing the set within a small volume.

1,551,391—E. F. Hennelly, filed August 29, 1921, issued August 25, 1925. Assigned to General Electric Company.

Electron Device of high power construction where the plate electrode is positively spaced within the evacuated container by compressible holder elements which are secured to the surface of the cylindrical plate electrode and frictionally engage the inner walls of the tube for resisting displacement of the anode under conditions of use for handling.

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The Institute of Radio Engineers

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M. B. Sleeper	L. B. Bender
A. H. Taylor	S. Kruse
E. L. Chaffee	P. O. Pederson
Leslie McMichael	G. Ferrie
A. Meissner	Fernando Ramirez
E. Yokoyama	Mario de Barros Barreto

Committee on Standardization

Ralph Bown (Chairman)
Lloyd Espenchied—H. W. Nichols
L. A. Hazeltine
R. H. Langley
A. E. Reoch
C. H. Taylor
L. E. Whittemore

Committee on Revision of the Constitution

J. V. L. Hogan (Chairman)
Melville Eastham
Lloyd Espenschied

Committee on Advertising

W. F. Hubley
Donald McNicol

INSTITUTE ACTIVITIES

Institute News and Notes

The first Fall meeting of the Section was held on the evening of October 8. In the next issue of the PROCEEDINGS it is hoped we shall be able to announce the meetings schedule for the Winter months. Members visiting San Francisco should plan to be present at these meetings when convenient. Meetings are held on the second Thursday of each month at the Engineers' Club.

Los Angeles Section

Preliminary work is now under way looking to the organization of a Section of THE INSTITUTE at Los Angeles, California. Col. J. F. Dillon, of the San Francisco Section, is representing headquarters in the organization plans.

Standards Committee

At the October 14 meeting in New York of the Standards Committee, the following members of the committee were in attendance: Ralph H. Bown, chairman, R. H. Marriott, J. V. L. Hogan, Donald McNicol, C. B. Joliffe, L. E. Whittemore, L. A. Hazeltine, E. H. Armstrong and H. M. Turner. The published Report should be ready for distribution to members early in the coming year.

Chicago Section

The Chicago Section of THE INSTITUTE held a meeting in the rooms of the Western Society of Engineers, Chicago, on the evening of October 9. Hugh A. Brown, assistant Professor of Electrical Engineering, University of Illinois, presented a paper on "Voice-Controlled Transmission in Radio Telephony."

Inaugural Meeting in Toronto

The new Canadian Section of THE INSTITUTE, organized in Toronto, Ontario, held an inaugural banquet and get-together meeting on the evening of October 2. The banquet was served in the dining-room of the Ward St. works of the Canadian General Electric Company. Eighty radio engineers and workers attended the meeting. Among those present were: Prof. T. R.

Rosebrugh, Prof. W. H. Price, H. C. Don Carlos, C. L. Richardson, W. J. Hevey, Keith Russel, C. S. Mallett, Prof. A. R. Zimmer, Prof. Pacent, J. J. Ashworth, F. A. Gaby, O. E. Forrest, A. S. Edgar, W. L. Amos, T. Rogers, H. Swift, W. B. Cartmel. The meeting was attended also by Donald McNicol, vice-president, New York, and by C. P. Edwards, Director of Radio Service, Department of Marine and Fisheries, Ottawa, Canada.

Washington Section

A meeting of the Washington Section of THE INSTITUTE was held in the Department of Commerce Building, Washington, on the evening of October 6. The annual election of officers took place, and a talk was delivered by Chairman A. Hoyt Taylor, on the subject: "High Frequency Phenomena."

Washington Conference

THE INSTITUTE's representatives appointed to attend the Radio Conference held in Washington, November 9, under the auspices of Secretary Hoover, were: R. H. Marriott, J. V. L. Hogan and Donald McNicol.

Membership Elections

At a recent meeting of the Board of Direction the following transfers were approved: To Fellow, Arthur F. Van Dyck. To Member, Austin Bailey, Powel Crosley, Jr., R. D. Duncan, Jr., E. H. Felix, C. M. Jansky, Jr., E. T. Jones, J. Marsten, E. L. Nelson, Arthur Nilson, R. S. Ohl, J. M. Thorburn, H. M. Wilby.

At the same meeting the following direct elections to grade of Member were approved: A. S. Blatterman, C. H. Burkhead, W. B. Cartmel, E. M. Deloraine, L. B. Henson, C. C. Jackson, H. L. Kirke, H. Levinson, L. H. Mansell, James Nelson, R. R. Pecorini, A. W. Peterson, R. J. M. Raven-Hart, F. Reichmann, E. R. Shute.

Applications Committee

The Applications Committee, of which Prof. J. H. Morecroft is chairman, holds regular monthly meetings, at which all applications in hand at the time are acted upon.

Annual Meeting in New York

President Dellinger, at the September meeting of the Board of Direction, appointed a committee to study the proposal for an annual meeting of THE INSTITUTE. Mr. J. V. L. Hogan is chairman. Report and decision in the matter shall be announced in an early issue of the PROCEEDINGS.

Membership Committee

The Membership Committee, of which Mr. L. E. Whittemore is chairman, held a meeting on October 21. Each member of THE INSTITUTE will shortly receive a circular on the subject, accompanied by an application form and postal card form, which it is hoped will receive personal attention to the end that all competent radio workers may be given an opportunity to apply for membership in THE INSTITUTE.

Institute Award

At the October meeting of the Board of Direction the Liebmann Memorial Prize for the year 1925 was awarded to Mr. Frank Conrad in recognition of the value of research work in short wave signaling carried on by him. This prize is awarded annually by THE INSTITUTE, and amounts to five hundred dollars in cash. The money award is accompanied by an appropriate letter of transmittal. Other radio engineers who have received this prize in past years are: R. A. Weagant, R. A. Heising, L. F. Fuller, C. S. Franklin, H. H. Beverage, and John R. Carson.

Radio Historical Museum

In the past year or two there has been occasional discussion on the subject of a historical museum into which might be gathered the various private collections of early relics of the art, large or small. The subject is one which should receive attention in the near future so that something may be done before the scattered, individual collections are dissipated. Members of the INSTITUTE who possess radio relics, or who know of the location of collections, are invited to forward information relative thereto to INSTITUTE headquarters. After an inventory has been made of available collections, a committee will be appointed by the President to take up the subject and see what can be provided in the way of suitable housing.

Technical Papers in Pamphlet form

Until such time as the PROCEEDINGS of the Institute are published monthly, instead of bi-monthly, as at present, some papers will be printed in pamphlet form in addition to those which appear in the PROCEEDINGS. One paper now available is "Recent Advances in Marine Radio Communication," by T. M. Stevens, of the Radio Corporation of America, New York. Copies of this paper may be obtained by writing to the Secretary of the INSTITUTE, 37 West 39th St., New York.

AN INVESTIGATION OF TRANSMISSION ON THE HIGHER RADIO FREQUENCIES*

By

A. Hoyt Taylor

(Superintendent of Radio, Naval Research Laboratory, Anacostia, District of Columbia)

The object of this paper is to present graphically and in systematic form information which is a summary of the range data for various frequencies so far as it can be estimated from the extensive experiments carried on by the Naval Research Laboratory at Anacostia, supplemented by considerable information from outside sources which has come in various ways to our knowledge.

Two further objects of this paper are, first, to indicate the regions which require further exploration, and second, to show the places where certain transition phenomena of a more or less abrupt character occur as the frequencies are varied from 100 kilocycles to 20,000 kilocycles. This should bring out the peculiarities of high frequency transmission and serve as a guide, in a general way, in formulating policies looking forward to the possible wider adoption of high frequency communication.

The attached range chart is based on the following considerations:

(a) 5 kilowatts in the antenna.

(b) Average antenna installation.

(c) Communication between points on the same meridian.

The chart is nevertheless generally applicable to east-and-west communication or any communication where there is considerable time difference between the points involved, provided due accounts are taken of this time difference. Nevertheless it must be admitted at the start that the problem is much more complicated for such a condition, especially where there is a very large number of hours of time difference between two points.

There is, of course, considerable difference between the daylight ranges, summer and winter, but not anything like the differences which occur in the night range. Therefore the line on

*Received by the Editor May 9, 1925.

the chart indicating daylight ranges must be considered as giving average ranges, summer and winter, but for the night range, the lower dotted line indicates the winter night range and the upper dotted line indicates summer night range.

A cross entered on the line indicates the limit of actual exploration and the extension of the line beyond the cross indicates the probable ranges. A "U" entered on the line, indicates, if bounded on the right and left by two arrows, a region (generally a short region) within which communication is uncertain—or in case the "U" stands further out to the right hand side of the diagram, it indicates that for ranges longer than those corresponding to the position of the "U," communication begins to become uncertain.

As an illustration, take the 4,000 kilocycle band. The daylight communication is set at about 750 miles. The summer night communication at about 7,000, but uncertain after 3,000. The winter night communication extends to 10,000 miles, but is subject to some uncertainty after 6,000. Another instance of the use of the "U" is shown in the 6,000 kilocycle band, in which there is an uncertain region between 150 and 400 miles beyond which the range again becomes certain and extends to 10,000 miles (probably), but is uncertain after 7,000. That is, this frequency has two regions of uncertainty: one at close regions and the other at very distant regions. The use of the question mark ("?") is for the purpose of indicating unexplored regions. The use of the "M" on the diagram means that the radiation "skips over" or "misses" entirely the region indicated; therefore the "M" is always bounded by arrows right and left. An instance of this is the 15,000 kilocycle band which skips over the region from 75 miles to 700.

Starting with 100 kilocycles, the ranges of which are fairly well known from the performance of certain ship transmitters, we see that the daylight range is about 1,200 miles, the summer night range 2,000 but uncertain after 800 on account of heavy strays in the summer time, and the winter night range extends to 2,500, but becomes uncertain after 2,000 on account of the strays and fading. At 200 kilocycles, we find the daylight range shortened to 800 miles, the summer night range good on the whole for a greater distance than the 100 kilocycles. This is true because there are less strays at 200 kilocycles in the summer time than there are at 100. The winter night range, however, overlaps that for 100 kilocycles, going to 3,000 miles, altho uncertain after 1,800 on account of fading. At 500 kilocycles the

daylight range is still further shortened and the summer night range is not certain for any greater distance than the daylight range, but the winter night range is certain for much greater distance and the extreme winter night limit (2,000 miles) is more than three times the normal daylight range. At 1,000 kilocycles we see the daylight range still further shortened, but the night range considerably exceeding it even in the winter, whereas in the winter the extreme ranges are very greatly in excess of the normal daylight range, with, however, about half of the winter night range in the region of uncertainty. It should be stated at this point that the table is based entirely on continuous wave telegraphic communication. At 2,000 kilocycles the daylight range is cut to 125 miles and the summer night range is not a great deal better; but the winter night range is enormously greater than the daylight range; with, however, a great region of uncertainty in the winter night range, due to fading. The performance here is based on the Fleet's report of certain transmitters built by this Laboratory, and tested on shipboard. It is also based on amateur data to a certain extent.

Between 2,000 and 3,000 kilocycles, the phenomena show a rather abrupt change. At 3,000 kilocycles the daylight range is much greater than at 2,000, which is a reversal of ordinary behavior at lower frequencies. The summer night range is enormously extended and the winter night range still more so. We see that the reliable night ranges for summer jump to 2,000 miles, and as a matter of fact this figure is probably considerably underestimated. It is, however, desired to make the chart conservative at least in its application to the higher frequencies. The night ranges in the winter time, however, are certain up to 6,000 miles. Comparing 3,000 kilocycles with 100 kilocycles, we find the 100 kilocycles excels in daylight range, but that the 3,000 kilocycles greatly excels in possible night ranges. At 4,000 kilocycles we see the daylight range extended to 750 miles, the summer night range certain to 3,000, with a possible night range to 7,000; while the winter night range is certain to 6,000 with possible ranges to 10,000. At 6,000 kilocycles the daylight range extends to 1,000 miles and the summer night range to 4,000 with possible ranges to 7,000, whereas the winter night range probably goes to 10,000, but has not been fully explored. We may consider it uncertain at least after 7,000.

A new and interesting phenomenon makes its presence felt for the first time in the diagram regarding an uncertain period within short distances during the winter night transmission

namely, between 150 and 400 miles. In the next line on the diagram for 7,500 kilocycles, the daylight range has been further extended to 1,200 miles, thus nearly equalling the 100 kilocycle transmitter, but an uncertain region not far from the transmitter has been introduced between 100 and 350 miles during the summer night range and a skip, or entirely-missed, region, occurs in winter night ranges between 100 and 350 miles. This frequency has been explored to 4,000 miles, but it is believed that it will carry very much further. We may say that it is uncertain, however, after 8,000 and probably will carry to 10,000 on the winter nights. At 10,000 kilocycles we see that the "jump" or "miss" occurs both summer and winter and also by daylight, but the daylight jump is only about 500 miles, whereas the summer night jump is very great indeed. Very little exploration has been made of this frequency for nocturnal transmission, but there is good reason to suspect that the summer night jump is in the neighborhood of 2,000 miles and the winter night jump possible 4,000 miles. There is also good reason to believe that frequencies not far from 10,000 kilocycles can be successfully used for extreme night ranges. (In this connection, there should be noted Samoa's successful reception and practically steady intensity of tests from Schenectady, New York, on tests of 35 meters during night hours.)

Fifteen thousand kilocycles: Here the daylight jump is increased to between 600 and 700 miles and an uncertain region follows this to 1,000 miles, but beyond this, as far as the exploration has gone (4,000 miles), results are excellent. It is impossible to say what the daylight range will come to beyond 4,000 miles. It is known, however, that the missing region or the jump in this frequency is very great at night, both summer and winter. Very little exploration has been made here, but there are some data indicating that this frequency can be successfully used at 10,000 miles even at night. This statement is based on the establishment of two-way communication with Australian 2-CM (Sydney). He used 21 meters and this station used 20.8 meters, between 1 A. M. and 2.30 A. M. Eastern Standard Time. It was broad daylight in Sydney when the test commenced. It should be noted at this point that this Laboratory has so far not used more than 750 watts in the antenna in the twenty-meter band. Australian 2-CM was using still lower power. It is evident, then, that figures estimated for 5 kw. in the antenna, but actually based on experiments with less than 1 kw., ought to be fairly conservative.

At 20,000 kilocycles the exploration is very scanty indeed, but there is information at hand giving the distance of the daylight jump as in the neighborhood of 1,500 miles with an uncertain region extending to 2,500 miles and which probably is followed by a certain region for a considerable distance beyond. Absolutely nothing, however, is known of night ranges on these frequencies. The ranges of the direct, or earth-bound components are well enough known; they are about 60 to 70 miles for 15,000 kilocycles and in the neighborhood of 40 to 50 miles for 20,000 kilocycles.

It is not the purpose of this particular paper to go into details of the vast amount of information upon which the range chart is based, nor to divulge at this particular time the theory which is gradually forming in the minds of the engineers of this Laboratory, which we believe will account for these curious effects. The purpose is rather to serve as a practical guide to indicate what ranges may be covered at different frequencies and what ranges remain to be explored, and what we hope to get in the un-explored regions.

It may be stated at the present time, however, that some of the most valuable information confirming earlier data on the matter of the "skip" or "miss" region was obtained from the daily reports made by Major J. O. Mauborgne, United States Army, who took observations from an Army Transport, on 16, 32, 20.8, and 41.7 meters, all the way from New York to Panama. It is believed that the information concerning the uncertain regions and "skip" regions is fairly definitely known for daylight work. If any special criticism could be made of the range chart it would be that it underestimates the summer night ranges on 3,000 kilocycles.

When one considers the chart as a whole, the high frequencies show clearly their enormous superiority from a point of view of economy on power consumption and general cost, and further it is possible to obtain ranges with high frequencies which we cannot hope to equal with almost any practical amount of power on lower frequencies.

To apply the chart to east-and-west communication, we must, for the present, consider that during the hours when daylight obtains over the entire stretch, we apply the daylight range data. For the hours which night obtains over the entire stretch, we apply the night data. In the intermediate hours when part is sunlight and part dark, much further exploration will have to be made, but we do know that a sort of compromise condition

does exist and it does appear further that a 5 kw. transmitter, equipped with about four frequencies, would be in a position to obtain highly creditable ranges at any time of either day or night, and whether for north-and-south communication, or for east-and-west. We must, however, at present, until further information comes in from stations like Samoa, Guam, and Cavite, be forced to believe that the east-and-west problem is more difficult of practical solution.

It must be understood in referring to the range chart that estimates on range and references to the "missing" of "skipped" areas on the higher frequencies refer in the case of the daylight ranges to conditions existing in the middle of the day, and for the night ranges, to conditions existing in the middle of the night. For west-and-east work this must be interpreted as meaning conditions when the sun is half-way between the two meridians under consideration. It is well known, of course, that there is a more or less gradual transition from daylight to dark conditions; in fact it is not nearly as abrupt as one would anticipate it to be, especially in the summer time.

Since the first part of this paper was written, Samoa has reported successful reception of the 20.8 meter wave from the Naval Research Laboratory as early as 8 P. M. zone-plus-five time, which means that 6,000 out of the 7,000 miles between Washington and Samoa were traversed in daylight. Also it is well known that 20-meter signals from amateurs on the West Coast are now received as late as midnight or 1 A.M. zone-plus-five time, which could not have been done during winter nights. This is interesting as showing a gradual change in the skipped region for 20 meters. The skipped region is less in the daytime, gradually increases to somewhat less than 2,000 miles in the summer nights and very likely is considerably in excess of this in the winter nights, altho it is not known with certainty whether it ever comes down to earth again in the winter nights. One must therefore, conceive of the skipped distance on the higher frequencies undergoing a lengthening process as the night wears on, followed by a shortening process as daylight approaches. Most of the information on the higher frequencies must, of course, be considered as incomplete subject to future revision. Nevertheless certain fundamental things in the behavior of these frequencies seem to be quite definitely established.

Balboa also reports satisfactory reception of the 20.8-meter wave thruout the 24 hours at this time of the year, but it is not anticipated that this will be possible in the winter

time. Very likely the signals as received at Balboa will fade out during 6 or 7 hours during the winter nights. At the present time the 20.8-meter wave, with less than 1 kw. in the antenna, is more satisfactory for handling traffic with Balboa than Annapolis on 17,000 meters.

April 21, 1925.

SUMMARY: A preliminary range chart has been constructed for telegraphic communication, 5 kw. in the antenna, at various frequencies. The conclusions upon which the range chart is based are derived from experiments made by the Naval Research Laboratory, from experiments made by amateurs, and upon such data as the Laboratory has had access to from commercial and Government sources at home and abroad.

An attempt has been made to indicate in a general way the advantages and disadvantages of high frequency telegraphic transmission. Various critical regions are pointed out where new phenomena make their appearance: in particular these regions are (1) the region between 2,000 and 3,000 kilocycles where daylight ranges begin to increase with increasing frequency at the same time that the night ranges show extremely great increase and a degree of reliability which would be wholly unanticipated from obervations made at frequencies lower than 2,000 kilocycles. (2) A region around 6,000 kilocycles where an uncertainty develops during the winter nights at relatively very short ranges. (3) The development at successively higher frequencies of this uncertain range into the missing region which is most pronounced in the winter nights, and finally as the frequency is increased makes itself felt in the summer nights and at still higher frequencies even in the daytime.

The development of this missing region to extensive areas is shown to take place with frequency rise to 20,000 kilocycles. The chart also attempts to indicate, in a general way, the region of uncertain communication and the regions where further exploration is urgently needed. It is quite evident that the range data are far from complete and that many individual cases will be found in contradiction to the chart, but it does represent a sort of general average of the situation as it presents itself to the Engineers in the Naval service.

It is hoped that the publication of this data will promote useful discussion and collaboration in this new and interesting field. The data would no doubt have to be modified materially to make it apply to any highly directive system of transmission.

A NEW DIRECTIONAL RECEIVING SYSTEM*

By

H. T. Friis

(Bell Telephone Laboratories, Incorporated, New York)

Introduction

Reduction of static interference, or to state it more correctly, reduction of the ratio of static to signal, has been, almost since the beginning of the radio art, the most important problem in radio engineering. It is now well known that static disturbances have definite points of origin and that the impulses which are detected at a receiving station have definite directions of propagation. A receiving system having no directional selectivity is, therefore, affected by static impulses from all directions and, in spite of many inventions, it has not yet been possible to improve its signal-static ratio except by limiting the frequency band transmitted. A system which, however, is so designed as freely to receive waves arriving from a limited range of directions is susceptible only to static disturbances propagated within that range, and large improvements in signal static ratio have been claimed for different types of directive antenna systems during the past few years.[1]

A directional receiving system for radio telephony in which directional selectivity is obtained by combining the output voltages from two antennas is described in this paper. The main feature of the system is the arrangement for controlling the output voltages of the antennas, so that they may be combined to neutralize each other or to reinforce each other as desired. A double detection (super-heterodyne) receiver is employed and the output voltages, which are combined so as to produce the directional characteristic, are the intermediate frequency currents due to the waves received by the antennas and the beating oscillator currents. The control of these output voltages is effected by operating upon the beating oscillator currents.

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[1] See "The Wave Antenna," by Beverage, Rice and Kellogg, "Journal of the American Institute of Electrical Engineers. This paper contains an excellent list of all publications on directional systems up to date.

Requirements of Directive Antenna Systems

The impulsive nature of most static makes it impossible to discriminate against it by frequency selective means, the effect of the impulses being to set up in the selective circuits free oscillations having a frequency and damping determined by the natural periods and the time constants of the circuits. It follows, then, that static effects in a plurality of antennas can be made to neutralize each other only if the *natural periods* and *time constants* of the antenna circuits are alike. It may be emphasized that this paper deals with reception of speech signals which require a much wider frequency band of the receiving set than telegraph signals. In a receiving set for telegraph reception, which responds only to a very narrow frequency band, perhaps a few hundred cycles, it is not so important that the antenna circuits be exactly alike.

In a multi-antenna system, it is necessary to control the phase and intensity of the currents delivered from each independent antenna circuit, in order that their mutual reinforcement or neutralization may be properly accomplished. Hitherto this has been done by incorporating phase and amplitude control apparatus directly in the antenna circuits, and the addition of such apparatus has made it extremely difficult to achieve the requisite equality of the antenna circuits. For this reason, the use of directional receiving systems has been very limited and has been almost entirely restricted to the long wave signaling systems. Alexanderson[2] in his paper on the Barrage receiver recommends always making the antenna circuits aperiodic, but such high resistance circuits are very inefficient, and they will increase the "set noise"[3] very much.

By placing the means for controlling the output voltages in the beating current circuit which is substantially independent of the antenna circuits, these may be reduced to a simpler form and may readily be made accurately alike and more efficient, thereby making it possible to receive waves of any radio frequency with good directional selectivity.

It should be observed that directional discrimination against continuous waves of frequencies, different from the frequency band to which the antenna circuits are tuned, is not necessary as the frequency selection effected by these circuits and the inter-

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[2] Proceedings of The Institute of Radio Engineers, volume 7, page 363, August, 1919.

[3] The "set noise" is defined as the field strength of the weakest signal that can be received by the set when there is no static, and is determined by the amplifier noises.

mediate frequency circuits is sufficient to suppress all interference from such waves. Further, since the effect of static disturbances is to produce free oscillations in the antenna circuits at the frequencies to which the circuits are tuned, it is evident that the directional discrimination against continuous waves will be effective also against static impulses. This has been checked by experiments.

The receiving system of Figure 1 comprises two similar an-

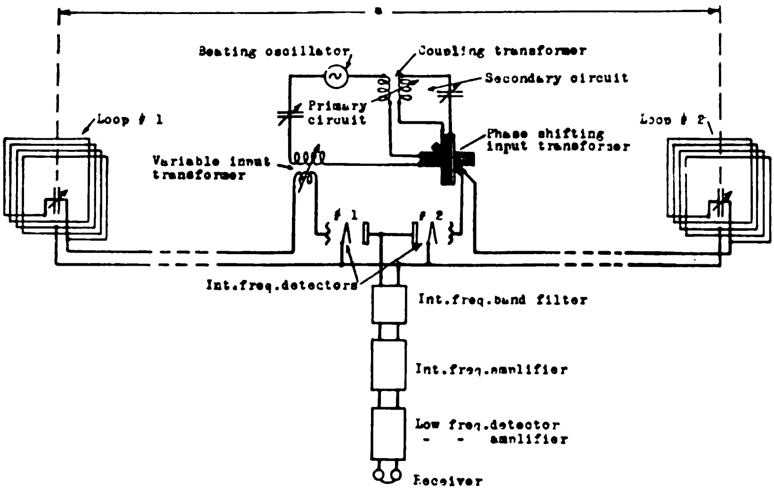


FIGURE 1—Schematic Circuit Diagram of "Two-Loop System"

tenna circuits, 1 and 2, the antennas (each of which is a loop) being so disposed as to produce in the two circuits emf's which are, in general, different in phase. The mid-points of the loops are connected to the "ground" of the set in order to reduce open antenna effects. The loops are arranged to have their planes substantially coincident and perpendicular both to the ground and to the plane of the desired wave. Their centers are spaced apart a distance a, and it will be shown later that this distance should preferably be one-twelfth of the wave-length it is desired to receive. An intermediate frequency detector tube is used for each antenna, the plate of these tubes being connected in parallel to the intermediate frequency filter. The secondary winding of a variable coupling transformer is inserted in series with the lead connecting the end of loop number 1 to the grid of its detector tube. This transformer supplies the beating oscillator current to the antenna circuit 1. The beating oscillator current is supplied to antenna circuit 2 thru a phase-shifting transformer which com-

prises two fixed coils having their planes mutually perpendicular and a third coil which can rotate inside the fixed coils. One of the fixed coils, the primary windings of the beating current input transformer to antenna circuit 1 and of a coupling transformer, and a tuning condenser make up a primary circuit tuned to the beating current frequency. The other fixed coil is part of a secondary tuned circuit which is loosely coupled to the primary circuit. At resonance, the currents in the primary and the secondary circuit are 90° out of phase and any desired phase angle of the beating current input to antenna circuit 2 can therefore be obtained by rotating the phase coil. The rest of the set, namely, the intermediate frequency filter mentioned above, the intermediate frequency amplifier, the low frequency detector and low frequency amplifier are well-known apparatus of the type usually employed in double detection sets and no detailed description is necessary.

The Operation of the System

It will be assumed that the system is intended to receive waves of frequency f propagated horizontally and in the direction from right to left in the diagram, Figure 1. The adjustments that should be given to the system and the preparations that should be adopted for the antenna spacing will be determined by considering the reception of horizontally propagated waves of the frequency f arriving in a direction at any angle β to the desired direction of reception (see Figure 2). As a measure of the directional effect will be taken the intensity of the resultant intermediate frequency current thru the intermediate frequency filter. The manner in which this varies with the direction of propagation, assuming a constant field intensity, defines the directive selectivity.

Two factors enter into the variation of the emf's produced

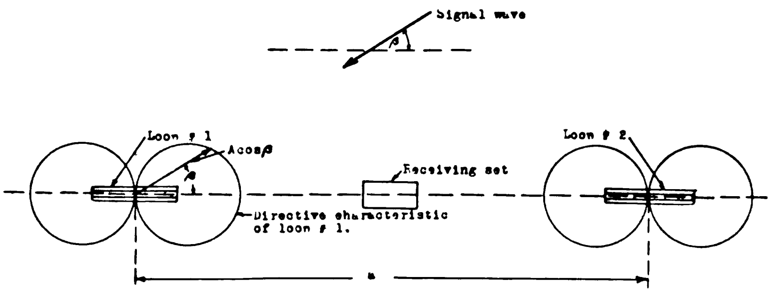


Figure 2—Top View of "Two-Loop System"

in the antenna circuits 1 and 2 by the waves. These are, first, the amplitude variation due to the directive properties of the individual loops, and second, the phase difference between the two emf's which depends upon the effective separation of the loops in the direction of propagation. The instantaneous values of the two emf's impressed upon the intermediate frequency detectors from the antennas may, therefore, be expressed by the equations:

$$e_1 = A_1 \cos\beta \cos 2\pi f t$$
$$e_2 = A_2 \cos\beta \cos\left(2\pi f t + \frac{2\pi a \cos\beta}{\lambda}\right) \qquad (1)$$

where A_1 and A_2 are the amplitudes corresponding to zero value of the angle β, that is, for the favored direction of propagation, and λ is the wave length corresponding to the frequency f. It is preferable to use two loops exactly alike, in which case the values A_1 and A_2 are the same. The subscripts 1 and 2 here, and in what follows, refer to the circuits 1 and 2, respectively.

From the beating oscillator there will also be impressed upon the detectors, emf's of a frequency f_o and of relative phases and amplitudes determined by the adjustments of the input transformers. The values of these emf's may be expressed by the equations:

$$e_1' = B_1 \cos 2\pi f_o t$$
$$e_2' = B_2 \cos(2\pi f_o t + \varphi) \qquad (2)$$

in which φ is the phase difference determined by the adjustment of the phase shifting input transformer, and B_1 and B_2 are the amplitudes of the emf's.

The important components of the currents in the output filter following the intermediate frequency detectors are those of the difference frequency $f - f_o$. These components are given by the equations:

$$i_1 = K_1 A_1 B_1 \cos\beta \cos 2\pi(f - f_o)t$$
$$i_2 = K_2 A_2 B_2 \cos\beta \cos\left[2\pi(f - f_o)t + \frac{2\pi a \cos\beta}{\lambda} - \varphi\right] \qquad (3)$$

the first of which represents the output current as a result of the intermodulation of the emf's e_1 and e_1' and the second of which represents the output currents due to the emf's e_2 and e_2'. The factors K_1 and K_2 involve the detecting efficiencies of the detectors and the total impedance of their output circuits. They will, in general, not be exactly alike but their ratio will be practically constant and they will not affect the relative phases of

the emf's, due to the fact that the two detectors have the same output circuit. The resultant effect in the signal reproducer is proportional to the sum of the two currents of equation (3). Since it is required that the two currents neutralize each other for certain directions of propagation, it is obviously necessary that their amplitudes must be made equal, that is

$$K_1 A_1 B_1 = K_2 A_2 B_2 = I \tag{4}$$

This may be done by varying the ratio of the beating current inputs B_1 and B_2 by means of the variable input transformer to antenna circuit 1. The amplitude of the resultant current is now given by the equation:

$$I_\beta = 2\, I \cos \beta \cos \left[\frac{1}{2} \left(\varphi - \frac{2\,\pi\, a \cos \beta}{\lambda} \right) \right] \tag{5}$$

It is evident that the resultant current will be zero under two conditions, first, when β equals $\frac{\pi}{2}$, that is, when the wave is propagated at right angles to the common plane of the two loops and second, when

$$\varphi = \frac{2\,\pi\, a \cos \beta}{\lambda} + \pi \tag{6}$$

One of the most useful adjustments, altho not necessarily the best under all circumstances, is that under which the signals from waves propagated in the opposite direction from the favored are neutralized. The angle β corresponding to this case is equal to π and the requisite phase displacement of the two beating current inputs is $\varphi = \pi - \frac{2\,\pi\, a}{\lambda}$[4]. The directive characteristic of a system having this adjustment is given by the equation

$$I_\beta = 2\, I \cos \beta \sin \left[\frac{\pi\, a}{\lambda} (1 + \cos \beta) \right] \tag{7}$$

which is derived directly from equation (5). The signal producing current corresponding to waves received in the favored direction is denoted by I_o, which is the value of I_β for the special case of $\beta = 0$ or

$$I_o = 2\, I \sin \frac{2\,\pi\, a}{\lambda} \tag{8}$$

Equation (8) indicates that the distance between the loops

[4] Note that φ is a function of wave length λ, which means that a complete balance can be obtained only for one frequency in the signal. The maximum relative changes of the frequencies in a speech signal will, however, always be so small that the amount of unbalance will be negligible.

should not exceed a quarter of the wave length $\left(\frac{2\pi a}{\lambda}=\frac{\pi}{2}\right)$. For small separations, the signal current I_o increases proportionally to the separation of the loops and reaches half the maximum possible value (or the same value as one loop alone would give) when the separation is one-twelfth of the wave length. For separations greater than this, the gain increases more slowly than the separation.

The ratio of amplitude of the current produced by a wave propagated in the direction β to that of the current produced by a wave of equal intensity propagated in the favored direction, is expressed by the equation

$$R = \cos\beta \, \frac{\sin\left[\frac{\pi a}{\lambda}(1+\cos\beta)\right]}{\sin\frac{2\pi a}{\lambda}} \tag{9}$$

As the separation of the loops is reduced, the ratio R approaches the limiting value

$$R' = \frac{\cos\beta}{2}(1+\cos\beta) \tag{10}$$

R' has maximum values for β equal to 0°, 120°, and 240° and has zero values at angles of 90°, 180°, and 270°. The ratio is naturally equal to unity when β is zero. The maximum ratios recurring at 120° and 240° are equal to 0.125. For one-quarter wave-length the ratio R is 0.1923 for angles of 120° and 240°.

The general form of the directive characteristic is shown in Figure 3, in which polar co-ordinates are used. The dotted figure represents the characteristic corresponding to an antenna separation of one-quarter wave length, and the full line that for the limiting condition which is closely approximated for all antenna spacings of less than one-twelfth wave lengths. It is evident that a distinct advantage in respect to the relative amounts of signal and interference received is to be gained by reducing the antenna separation to one-twelfth of a wave length, altho this is accompanied by a reduction of signal strength. To decrease the spacing still further will economize on the land required for the system, but will require, first, more amplification with a corresponding increase in "set noise" and, second, a higher degree of stability of the system.

It has already been pointed out that neutralization of the interference may be secured in other directions than that considered in the foregoing analysis, and it may happen in practice

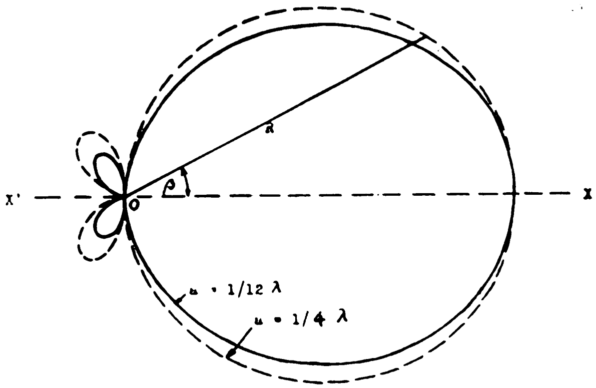


FIGURE 3.—Directive Characteristic of "Two Loop System"

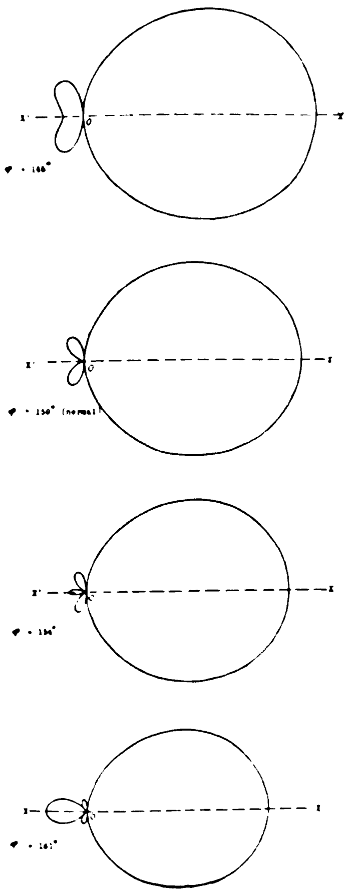


FIGURE 4.—Directive Characteristics of "Two-Loop System." (u = 1/12)

that the most troublesome interference may be that arriving from some other general direction. The simplicity of the phase control arrangements and the fact that the control can be exerted without disturbing the equality of the antenna circuit constants, make it a very simple matter to secure the most efficient suppression of interference under any circumstances. All that it is necessary to do is to adjust the secondary coils of the two beating current input transformers until the signal is heard with the least amount of interference.

The effect of changing the phase angle φ is shown in Figure 4, in which directive characteristics are calculated for φ equal to 145°, 150°, 154°, and 161°, all corresponding to a loop separation of one-twelfth of the wave length. The curves show that the setting of the phase angle is not very critical where it is desired to reduce fairly evenly distributed static interferenice.

EXPERIMENTAL VERIFICATION

Short Wave Two-Loop System

The system described in the foregoing was tried out at Cliffwood, New Jersey, with signals in the broadcast frequency range because the short wave length of these signals make it possible to mount the two loops on a construction similar to a turn-table which is capable of rotation in a horizontal plane. The photographs in Figures 5 and 6 show the general construction, and the receiving set itself is shown in the illustrations, Figures 7 and 8. The receiving set is installed in the 6×6-foot house, shown between the two bridges in Figure 5. The whole construction rotates around a bolt in the center of the house, the house being mounted on four 8-inch truck casters and the outside ends of the bridges carrying the loops being supported by ordinary wheelbarrow wheels. The distance between the loops is 34.4 meters, giving an upper limit of the wave lengths of signals to be received of approximately 600 meters. The loop diameter is 1.76 meters and each of the loops are wound with 6 turns of bare number 16 copper wire. One loop can be turned a few degrees as it is quite important that the loops point in the same direction.

The circuit is the same as that shown on Figure 1, except that the tuning condensers for the loops are mounted in the receiving set in order to improve the facility of tuning adjustments. The ends and mid-point of each loop are connected respectively to the tuning condensers and ground of the set by 3 wires suspended in a horizontal plane and 30 cm. apart.

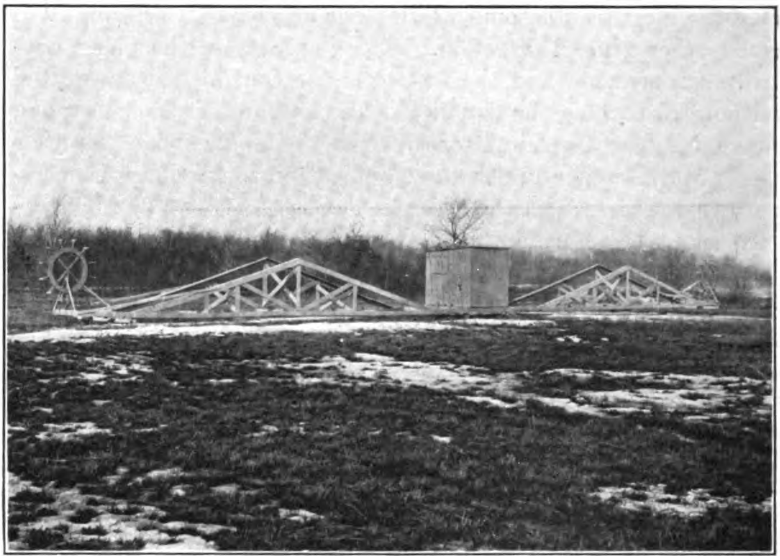


FIGURE 5—Short Wave Two-Loop System. Shows the Two Loops and the House Containing the Receiving Set

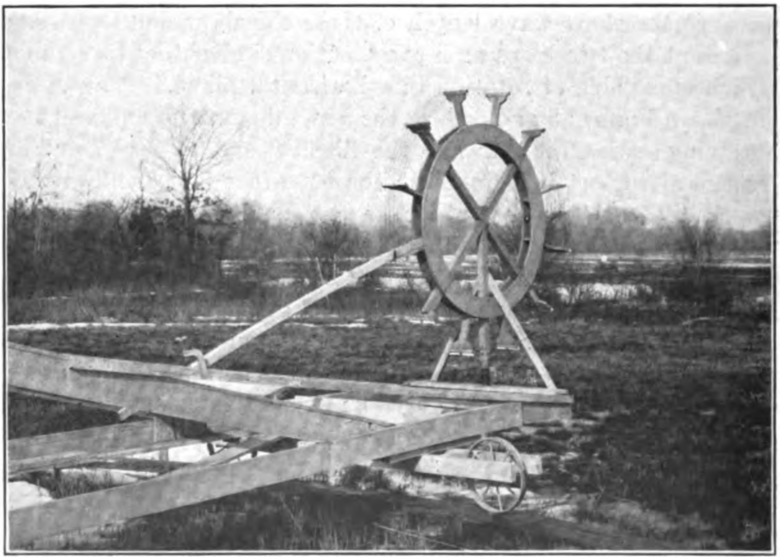


FIGURE 6—Short Wave Two-Loop System. One of the Loops

Figure 9 shows an experimental directional characteristic of this system. The incoming signal, in this case from station WEAF ($\lambda = 492$ meters), was first balanced out with the loops

pointing towards WEAF ($\beta = 180°$), and then the whole system was turned. The radius R in the diagram is proportional to the input voltage on the low frequency detector and there is seen to be a good agreement between the experimental curve and the theoretical curve shown in Figure 3.

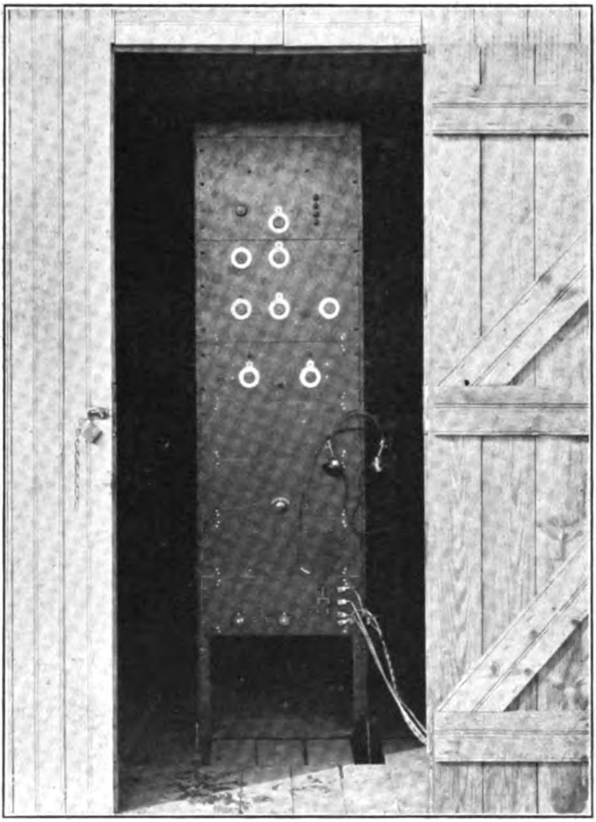


FIGURE 7—Short Wave Two-Loop System. Front View of Receiving Set

It may be pointed out that the tuning of the system to a signal is quite simple. First, one of the loops is short-circuited and the other loop tuned and the beating oscillator frequency adjusted as for an ordinary double detection receiver. Then the beating oscillator circuits are tuned up and finally the previously short-circuited loop is tuned. The set is now ready for the two adjustments of beating oscillator inputs giving a minimum of interference. It is quite convenient, especially at long wave

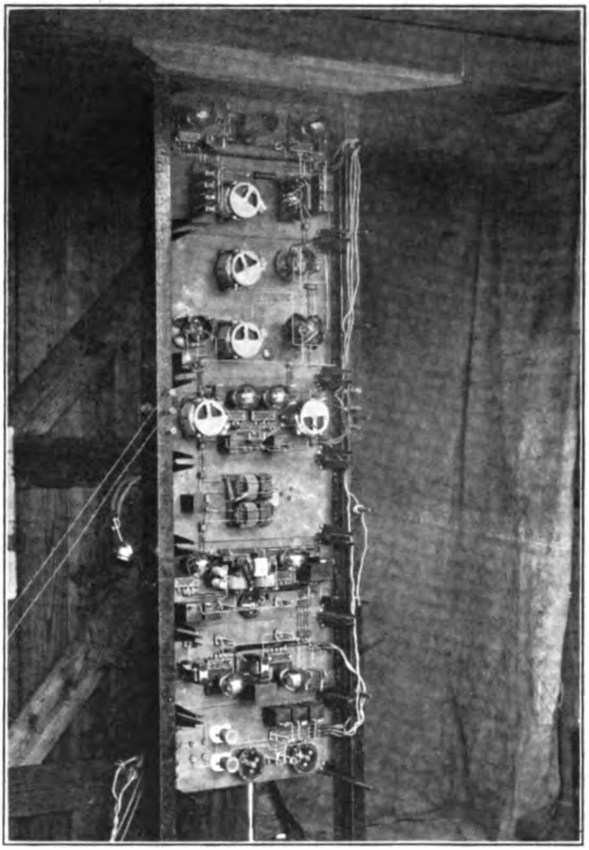


Figure 8—Short Wave Two-Loop System. Rear View of Receiving Set

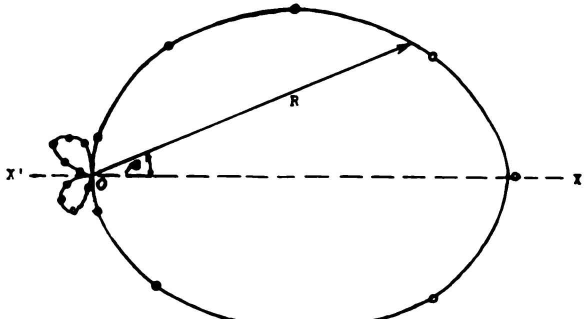


Figure 9—Experimental Directive Characteristic of "Two-Loop System"

lengths, to tune the set on a local oscillator, the frequency of which is adjusted to zero beat with the desired signal.

The short wave system described above was tested during the summer months, 1924, at Cliffwood, New Jersey, and found to verify all conclusions derived from the shape of its directional characteristic. The reduction in spark interference when receiving signals from broadcast stations in Philadelphia was especially noticeable. The spark interference at Cliffwood, New Jersey, is mainly due to stations along the shore and on ships around New York Harbor and the New Jersey coast, so that the interfering waves are coming from "behind" when the system is adjusted for receiving Philadelphia stations. On many occasions it was possible to reduce the summer static interference so much that talk from broadcasting stations which was absolutely unintelligible when received on one loop alone, was made clearly intelligible by the two-loop system. It may here be pointed out that static interference at broadcast frequencies, in the summer, is mainly due to local thunderstorms and it is therefore generally directive, but the direction is quite arbitrary. A gain in regard to reduction of static interference can, therefore, not always be expected since the static may come from the same direction as the signal wave. At long waves the direction from which static waves arrive is generally southwest, so that a considerable reduction in static may be expected when receiving signals from Europe.

Long Wave Two-Loop System

A receiving system suitable for reception of single sideband speech signals of 5,000-6,000 meters wave lengths was constructed and tested out during the fall of 1924 at Cliffwood, New Jersey. The locations of the receiving set and the two loops were permanent, as it is obviously not practicable to rotate a system 400 meters long like a turn-table. However, this limitation is not so important because long wave receivers are generally used for reception of signals from one direction only. The main object was to receive signals from Europe and the two loops were therefore located in a vertical plane at an angle of 60° east of north and 400 meters apart. It was found later on, however, that it is equally important, when the loops are located, to consider the direction from which the interfering waves arrive. If, for instance, the angle between the average direction of the interfering waves and the signal waves is only 90°, then quite an improvement can be obtained by making the angle β between the signal wave direction and the plane of the two loops 30° instead of 0°

(see diagram in Figure 10). During the winter months it was found that static interference was coming from the southwest during the night and from south-southeast during the day, in which case it is preferable to have two sets of loops and switch from one set to the other according to the direction of the static.

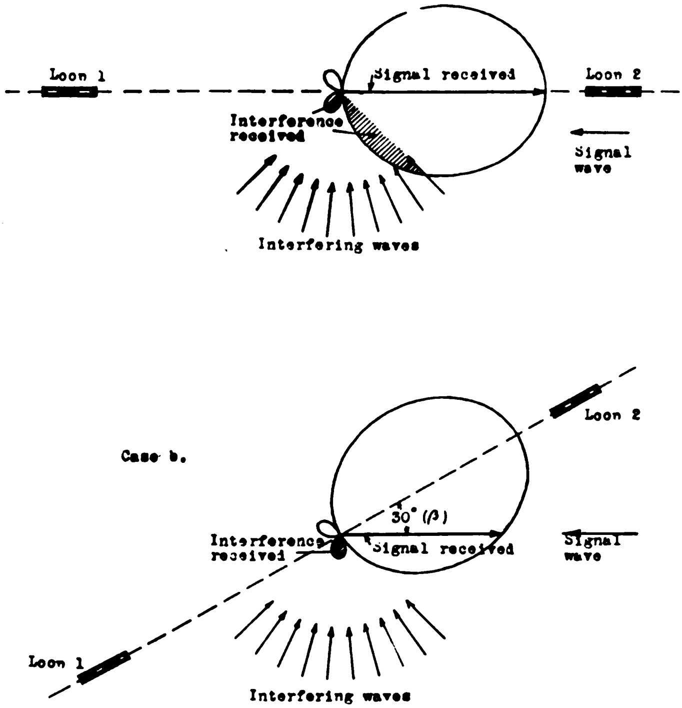


FIGURE 10—Effect of Changing Location of Loops. Location in Lower Figure Improves Signal-noise Ratio in Case of Distribution of Interference as Shown

The circuit diagram is the same as that shown on Figure 1, with the exception of the antenna circuits which are shown in Figure 11. Double tuned antenna circuits are used here at long waves, due to the fact that a low resistance single loop circuit is too selective for speech signal reception. The loop circuit and the secondary circuit are coupled together electromagnetically by means of a variable coupling coil. Electrostatic coupling was tried also, but it was found much more difficult to obtain a good balance of impulse interference with this kind of coupling. The reason for this is as follows: Let it be assumed that both the electrostatically and the electromagnetically coupled circuits are tuned, using weak coupling, to a certain frequency and that then

the couplings are increased in order to get the well-known resonance curves with two peaks. Now, in the case of electrostatic coupling the amplitudes of these two peaks will only be alike if the time constants are the same for the primary and the secondary circuit, or if the inductances and capacities have the same values in both circuits, while in the case of electromagnetic coupling the two peaks will always have the same amplitudes.

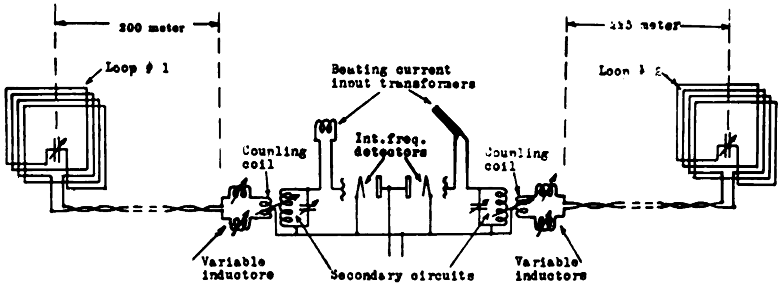


FIGURE 11—Antenna Circuits of "Long-Wave Two-Loop System"

The loops are installed in 10×10-foot wood houses in order to make the system independent of the weather conditions. Each loop is 8 feet square and has 40 turns of bare number 14 copper wire spaced ¾ inch. The loops can be rotated because it is quite important to measure the "loop minimum"[5] and also to have the two loops pointing in the same direction. The loop tuning condenser is mounted in the loop, but the loop circuit tune can be varied a little in the building where the set is installed by means of two variable inductances inserted in series with the two wires that connect the loop with the coil coupled to the secondary circuit. The arrangement of the antenna circuits shown in Figure 11 makes it possible to use an ordinary weatherproof pair of twisted wires lying on the ground for the 200-meter connection between the loop and the set without introducing appreciable losses in the antenna circuits. This is due to the low impedance of the variable inductors and the coupling coil terminating the line.

The illustration, Figure 12, shows one of the loop houses. The total weight of house and loop is less than 2,000 lbs., so that a team of horses can readily move the loop to any desired location. Such flexibility is very desirable in a "long wave two-loop system" which is still in its experimental stage.

[5] By the "loop minimum" is understood the ratio of maximum to minimum emf. induced in the loop as this is rotated. This ratio ought to be at least 25 times.

By placing the two loops close together it can be determined whether the constants of the two antenna circuits are sufficiently alike. For a small distance "a" between the loops, it should be possible to balance out signals from all directions because the emf's in the loops are then always in phase. The "balance," that is, the decrease in interference when receiving with the combined loops, as compared to the interference when receiving on one loop alone, depends only upon the constants of the two antenna circuits, and it ought to be at least 40 times. The full

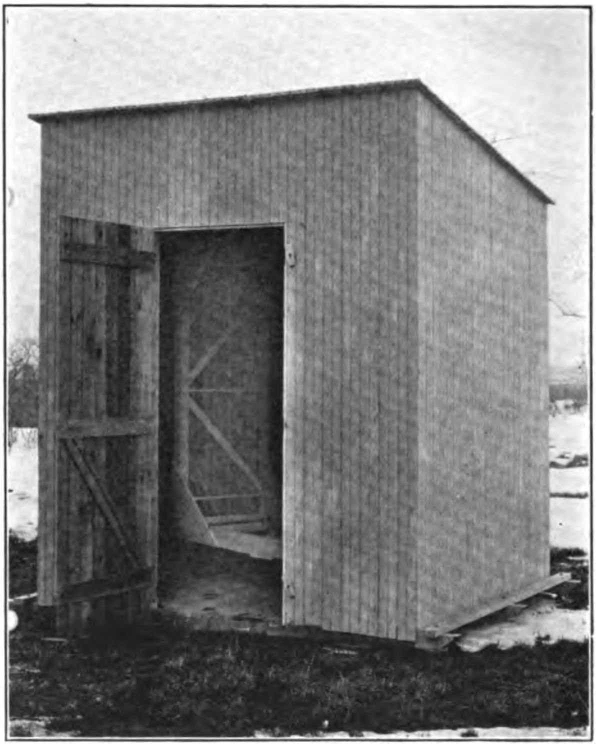


FIGURE 12—"Long Wave Two-Loop System." One of the Loop Houses

length of twisted wire is naturally used when experimenting with the loops close together, and while the loops must point in the same direction, this direction is chosen so that the coefficient of coupling between the loop circuits is at a minimum. After the antenna circuits are thus thoroly tested, the loops can be moved to their right location.

In order to compare this system with other directional systems it is necessary to measure its signal-to-noise ratio and com-

pare it with the signal-noise ratio of some standard antenna system. So far the loop antenna has been our "standard" for comparison. Such measurements require that both the selectivity and the "set-noise" be the same for the standard system and the system the improvement of which is to be determined. In the case of the two-loop system it is easy to satisfy these requirements because the standard system may be obtained by merely short-circuiting one of the loops. The signal amplitudes received by a single loop and by the two balanced loops are equal when the distance between the loops is one-twelfth of a wave length, in which case the improvement in signal-noise ratio is equal to the increase in the noise when one loop is short-circuited. It is thus seen that it is not necessary actually to receive signals in order to determine the improvement of the system. If the distance between the loops is not one-twelfth of a wave length, then a correction factor must be added. The best method of measuring the increase in noise when one of the loops is short-circuited is to decrease the beating oscillator input, at the moment the loop is short-circuited, to such a value that the noise level remains the same. The decrease in beating oscillator input measures the improvement of the system.

If the character of the interfering noise has not changed when one of the loops is short-circuited, it is not difficult for experienced operators to adjust the beating oscillator input for equality of noise level and check results. But often static interference is so variable that several measurements have to be taken. The measurements described above are generally checked by measurements which employ a local warbler signal[6] introduced in one of the loops, and adjusted until it can just be heard thru the noise. The improvement is then determined by the ratio of the warbler signal inputs corresponding to reception with one loop short-circuited and with two balanced loops, respectively.

It is desirable to measure the improvement in the signal-to-noise ratio of this long wave two-loop system. Such measurements would include continuous data on signal-to-noise ratios, strength of static, and direction of static. So far the system has been available only in the winter time, so that it has been impossible to make systematic measurements. However, valuable information has already been obtained and the results are, that, whenever the improvement was only 2 to 3 times, it was found

[6]Warbler signal oscillators and their use is described in a paper on "Radio Transmission Measurements," by Bown, Englund, and Friis, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 11, number 2.

that the general direction of static was south-southeast, that is, the static direction was at right angles to the plane of the two loops while static from south, southwest, and west was always reduced 4-6 times, often 8-10 times, and sometimes even 20 times. On one day only could no improvement be noticed, but static was that day coming from the northeast. It may be emphasized that a 7-times reduction in static corresponds to a 7^2 or approximately 50 times saving in power at the transmitting station.

Condenser Antenna—Loop System

The receiving set of the "long wave two-loop system" described above can obviously be used to combine the output voltages of other types of antenna. A system consisting of a condenser antenna and a loop will thus give the well-known cardioid directional characteristic shown in Figure 13, and as it is unidirectional, such a combination is especially suited for the determination of the direction of static interference. The long wave set was therefore equipped with switches so that the two loop circuits ordinarily used could be interchanged with a loop and condenser antenna both located close by the set. The illustration, Figure 14, shows the condenser antenna which is made up of two 5×5-meter frames covered with copper netting and spaced 15 cm. apart, and the loop used can be seen towards

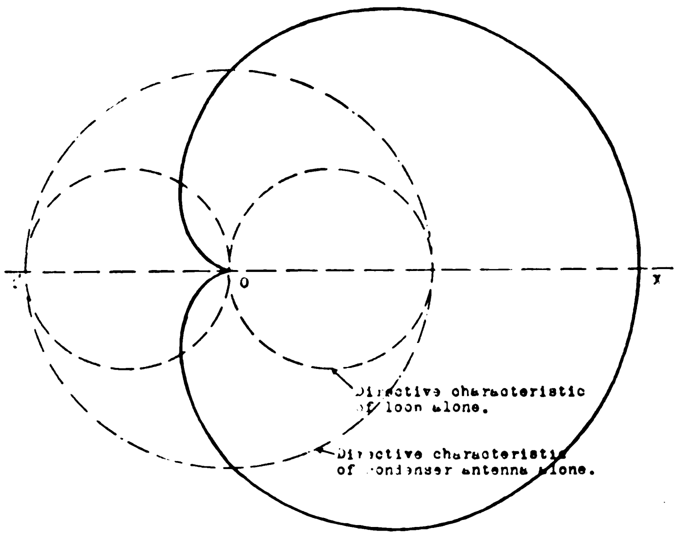


FIGURE 13—Directive Characteristic of "Condenser Antenna-Loop System"

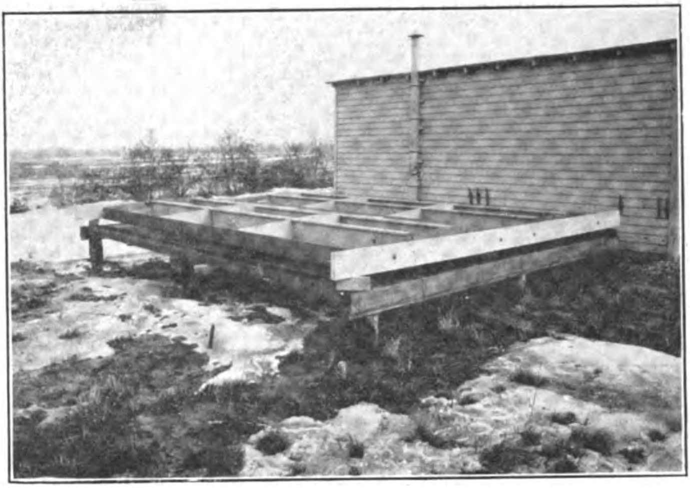


FIGURE 14—Condenser Antenna

left in the illustration, Figure 15, of the long wave receiving set. This system was used to determine the general directions of the static mentioned earlier in this paper and has naturally been of great help in the interpretation of the different improvement values obtained for the "two-loop system." So far the condenser antenna-loop system has shown that static at long waves in general has some directional characteristic. Sometimes the ratio of maximum to minimum static received when turning the loop was as high as fifteen times, that is, static was very directional.

The mathematical treatment necessary to derive the equation for a cardioid directional characteristic shown in Figure 13 is very simple and the procedure is similar to that used in the analysis of the two-loop system. The maximum and minimum signals are produced by waves propagated in the plane of the loop and they may be reversed by simply reversing the loop connections. Discrimination against interference from a specific direction may be secured by directing the plane of the loop towards the interfering source, the polarity of the loop being such that signals of that direction are suppressed. An experimental directive characteristic which was obtained by measuring the strengths of a continuous wave signal for different positions of the loop checked the cardioid curve in Figure 12 exactly.

The emf's induced by the signal in the loop and condenser antenna circuits are always in quadrature and the beating oscillator tuning circuits can therefore be made somewhat simpler than

FIGURE 15—Long Wave Two-Loop System. Front View of Receiving Set

shown in Figure 1 if the set is to be used only to produce the cardioid directional characteristic. Figure 16 shows how simple it is to change an ordinary double detection receiver for this purpose. The currents in the beating oscillator circuit 3 and the tuned circuit 4 are in quadrature and the two beating current emf's in the antenna circuits will therefore also be 90° out of phase. A small change in this phase angle may be obtained by detuning circuit 4 slightly and the relative values of emf's may be adjusted by varying the mutual coupling M. A good minimum

on spark signals and static, that is, a good balance of transient currents in two antenna circuits can only be obtained when the time constants of the two circuits are alike, and it is, therefore, in general, necessary to insert some extra resistance in the loop.

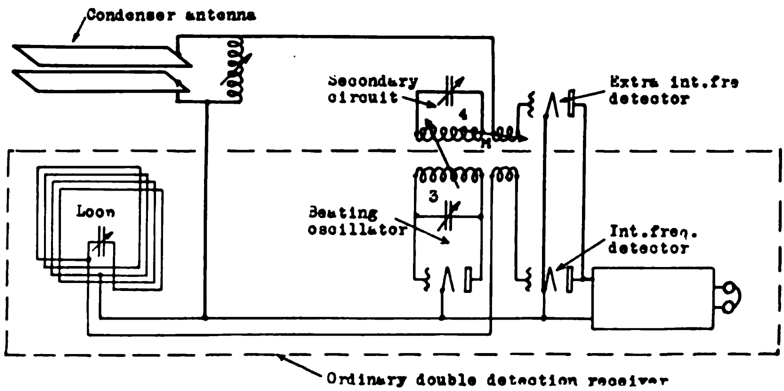


FIGURE 16—Schematic Circuit Diagram of "Condenser Antenna-Loop System"

The cardioid characteristic is inferior to the characteristic obtainable with a "two-loop system," but at long waves it has one great advantage, namely that it can be pointed towards any direction by simply turning the loop and it will give much more satisfactory results than a loop alone when it is desired to reduce interference from static or sparks, the direction of propagation of which makes an angle of at least 90° with the desired signal wave propagation.

CONCLUSIONS

A short wave directional two-loop system has been tested and shown to have the calculated directional properties. A similar long wave two-loop system has been developed and tested as far as possible under winter conditions of interference. Measured improvements in signal-to-static ratio range from unity to twenty, depending on the direction of the static. If this lies chiefly in the rear, the improvement is certainly 6 to 8 times. In fact, as far as experiments have gone, they indicate that, given a certain static distribution, the improvement could be calculated from the directional characteristic. Actual static distributions have, so far, favored the two-loop system as against the condenser antenna-loop system.

The directional antenna systems described in this paper re-

quire a double detection receiving set with a separate intermediate frequency detector for each antenna, and all necessary phase and amplitude adjustments are performed upon the beating current inputs. The system really makes use of the following general rule for double detection sets:

So far as the intermediate frequency currents are concerned, any change in phase, amplitude or frequency which it is desired to perform upon the antenna circuit current as a whole may just as well be performed upon the beating current.

The main advantages of this type of directional systems may be summarized as follows:

(1) Large reduction in static interference.

(2) Simple adjustments for interference reduction.

(3) Dimensions of not more than 1/12th of a wave length (for the two-loop system), so that only comparatively small areas are required.

(4) High efficiency antenna circuits so that excessive amplifications are not required.

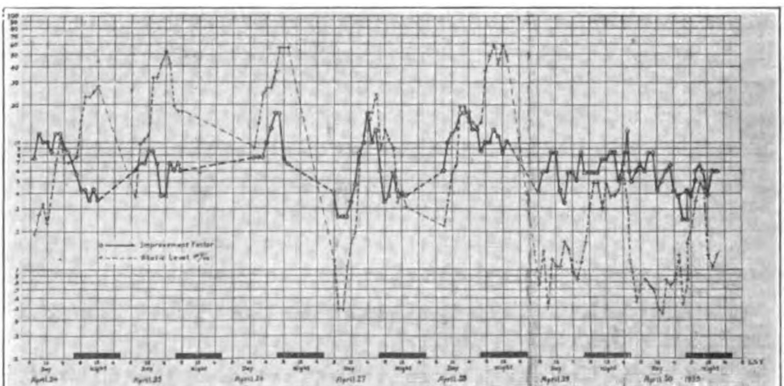
(5) Plenty of power is available in the beating current circuits, which simplifies the construction of the phase and amplitude controlling apparatus.

(6) The system can be checked quite readily.

(7) It is quite easy to house the system so as to protect it from the weather.

Appendix

Since this paper has been written it has been possible to take continuous measurements for one week on the long wave two-loop system described.



The results are given in the foregoing diagram. The full-drawn curve gives the improvement in signal-to-static ratio measured on a band approximately 2,500 cycles wide centering upon 57,000 cycles with the antenna located at Cliffwood, New Jersey, and directed for reception from England. The dotted curve gives the static level in $\mu V./m.$, as measured on one of the loops alone. The curves show that the improvement factor is approximately 10 times when the static level is larger than 5 $\mu V./m.$

SUMMARY: The paper discusses methods of combining the signal currents from the different antennas in a directional receiving system and a detailed description is given of a system by which all phase and amplitude adjustments are performed upon the beating current inputs of a double detection receiver. The theoretically derived shape of the directional characteristic of a two-loop system has been verified by experiments, and data on reduction of static for such a system are given.

AN ANALYSIS OF REGENERATIVE AMPLIFICATION*

By

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Present Conceptions of Regeneration

When Edwin H. Armstrong, in 1912, brought the phenomena of regeneration to the attention of the radio world, he undoubtedly gave rise to one of the greatest factors in the advance of radio today. Before regeneration, the electron tube served merely as a simple rectifier and amplifier. After regeneration, the same tube served as an amplifier of apparently unlimited extent. As is often the case in similar discoveries, the critical adjustments required and the enormous gain in signal strength so obtained gave the impression that the action was necessarily very complicated. Consequently at first very little attempt was made to explain the amplifying action except by stating that synchronous energy was fed from the plate circuit to the grid circuit which re-inforced the incoming signals.

*Received by the Editor, January 17 and February 28, 1925.

Practically all of the present discussions and analyses of regeneration can be grouped under two heads. First, that regeneration produces an equivalent reduction in the resistance of the grid circuit. Second, that regeneration is a voltage amplification due to the addition of the re-impressed grid voltage and the original applied voltage.

The first explanation of regeneration is the one more commonly accepted. This states that the equivalent resistance of the grid circuit decreases a definite (not proportional) amount as the tickler coupling is increased. A curve plotted between tickler coupling and the equivalent grid circuit resistance is a straight line and intersects the zero resistance axis at the point of critical regeneration. In this analysis an infinite amplification and an infinite response are obtained at critical regeneration regardless of the initial value of the signal voltage. This, of course, is impossible practically and so this type of analysis is always modified by the statement that the tube output is limited. This means that the characteristic curves are not straight lines in actual practice as we assumed in the theory. This variation from linear characteristics, which will be shown later to be the most important characteristic of regeneration, is absolutely omitted from almost all present mathematical analyses of these phenomena. We are, therefore, forced to reject the conclusions reached by such a method of analysis so far as they affect the conditions actually obtained in practice.

The second method of analysis shows that the final voltage applied on the grid is the sum of the initial applied voltage and the voltages introduced by the tickler. This analysis in some detail follows. When a regenerative circuit is brought near the point of oscillation, any voltage impulse on the grid will cause an oscillatory current which will die out at a rate depending on the resistance of the circuit and the value of tickler coupling. If the ratio of the amplitude of one cycle to that preceding is called "a" and if E_g is the amplitude of a given cycle, aE_g is the amplitude of the next. If an alternating voltage E_1 of resonant frequency is applied in series with the circuit, then the voltage on the grid for the first few cycles will be as follows:

$$E_1 \quad : \quad a\,E_1+E_1 \quad : \quad a\,(a\,E_1+E_1)+E_1 \quad : \quad - - - -$$

And for the n-th cycle will be $\dfrac{E_1(a^n-1)}{(a-1)}$

If "a" is less than "1" and "n" is infinite, then a^n is zero and

$$E_n = \frac{E_1}{(1-a)}$$

As the value of "a" approaches "1," E_n increases without limit. If "a" is larger than "1," the circuit will oscillate of its own accord. At critical regeneration $a=1$ and an infinite response is obtained regardless of the value of E_1. The criticism of the first type of analysis also holds here. That is, in practice, linear relations are not maintained, and there is no reason therefore to assume that any mathematically constructed analysis can be given as a correct explanation of regenerative phenomena.

If any further proof is needed to show the fallacy and inadequacy of the present analyses of regeneration, only a few facts are necessary. Assuming that either (or both) of the above methods of analysis are correct, it is evident that regenerative amplification is limited only by the ability of the operator to adjust the tickler to the critical value, and that with any given adjustment the final voltage will be directly proportional to the applied voltage. This means that the amplification is a constant. However, quoting from an article entitled "The Limit of Regeneration," by N. C. Little, in the August issue of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, for 1924, we find the following: "Results obtained with VT-1 and UV-201 tubes, both used with varying amounts of grid bias, show that *the relative magnitude of received signal in plate circuit to that impressed on the grid is inversely proportional to the latter.* This is called the inverse signal strength law. It states that the response of a system adjusted to critical regeneration is independent of the strength of the impressed signal, that no matter how weak the oscillating field surrounding an antenna may be, if the regeneration is pushed to its limit, a finite signal may always be obtained."

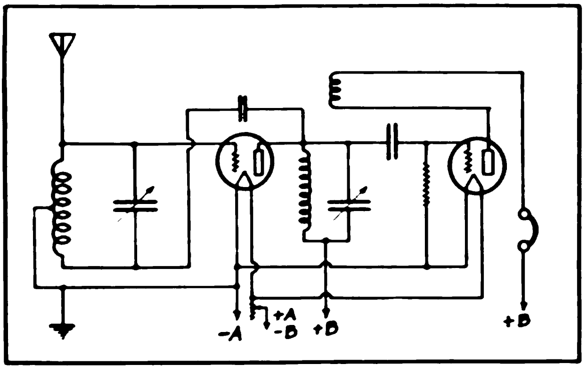
Obviously both of the above statements regarding the ratio of the final signal to the impressed signal cannot be correct. The first is quite evidently untrue, and the more experienced radio engineers will not accept the latter as an exact statement either. Something more than the previously advanced theories must be considered in the true explanation of regeneration and its limits.

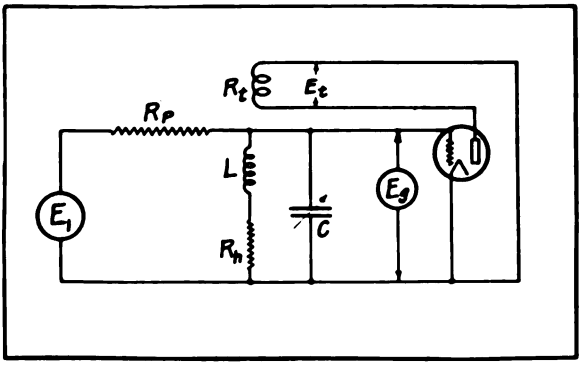
POWER BALANCE IN NON-REGENERATED CIRCUIT

To this end let us consider a few facts about the simple circuit shown in Figure 3. E_1 is an applied voltage, R_p the resistance equivalent to the tube impedance,[1] and R_h the resistance in the inductive leg of the tuned circuit of inductance L and capacity C.

[1] The action of a vacuum tube as an amplifier can be replaced by a fictitious generator of zero internal impedance, in series with a fixed resistance The voltage of this generator is μ (the amplification constant) times the voltage applied to the grid, and the resistance is equal to the plate impedance.

The resistance in the capacity branch is assumed so small as to be neglected.





FIGURES 1 and 2*

Let a voltage E_1 be impressed as shown. E_g will build up, reaching a final value which depends on the relative impedances of R_p and the tuned circuit. However, let us analyze the instantaneous conditions in the circuit. The power input at any instant is given by $E_1 \times I$. But $I = \frac{E_1 - E_g}{R_p}$ and therefore Watts Input $= E_1 \frac{(E_1 - E_g)}{R_p}$. The watts lost at any instant are given by the sum of the losses in the two resistances, R_p and R_h. The

* This figure shows the Rice method of balancing out capacity coupling between two tuned circuits. Divided plate circuit neutralization could be used with equal advantage.

loss in R_h is proportional to E_g^2. The watts lost are given by the sum of the following two terms:

Watts Lost $= \frac{(E_1-E_g)^2}{R_p} + \frac{E_g^2}{K R_h}$ where $\frac{E_g}{\sqrt{K}}$ is the effective voltage across R_h with E_g volts on the grid.

Before the final value of E_g is reached, the watts input will exceed the watts lost. This surplus is stored up in the tuned circuit as a charge on the condenser or in the magnetic field of the inductance. A stable condition is reached when the watts lost equals the watts input from E_1. Setting these two expressions equal and solving for E_g gives

$$E_g = \frac{E_1 K R_h}{R_p + K R_h}$$

This expression is the same as that obtained by assuming that E_1 divides between R_p and the tuned circuit in the ratio of their impedances.

The instantaneous conditions can be shown to a better advantage by means of curves. These curves are shown in Figure 4, and are plotted between E_g^2 and watts. The square of E_g is used as abscissa in order to make the loss curve for R_h a straight line. The advantage of this will be more forcibly shown later. Notice that the watts input is larger than the watts lost below the point of intersection. This point determines the final voltage and is called the point of power balance.

Power Balance in a Regenerated Circuit

In the above discussion, the only power available for building up the voltage E_g came from E_1. In a regenerative electron tube, however, the plate circuit introduces power into the grid circuit. If it introduces sufficient power, the tube will oscillate. In Figure 2, let E_1 be zero. Then in order to oscillate, the plate must supply the resistance losses in both R_h and R_p. In the regener-

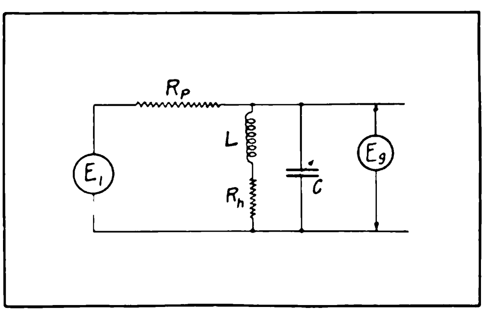


FIGURE 3

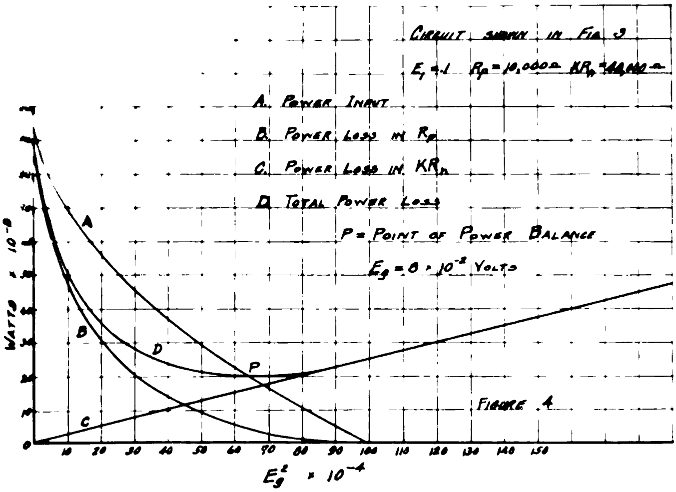


FIGURE 4

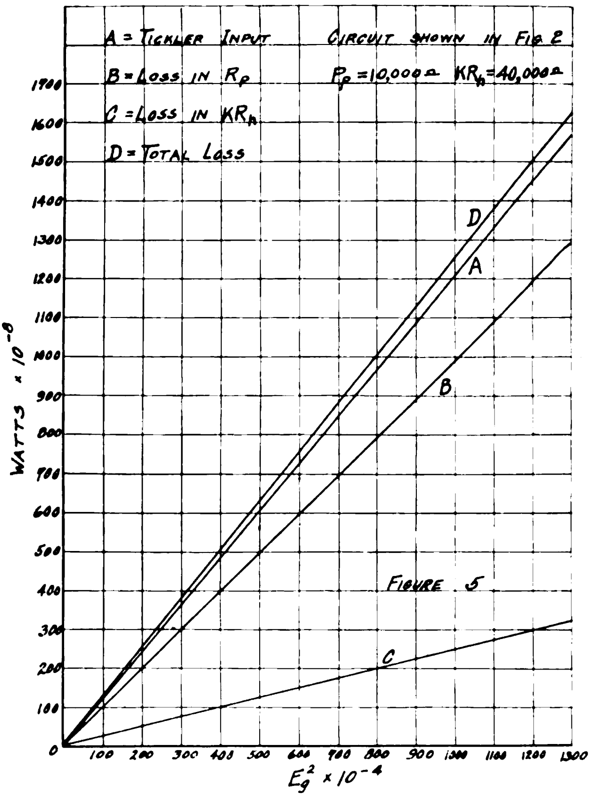


FIGURE 5

ative state, the power supply from the plate circuit for any value of voltage E_g is less than the losses in the circuit at the same value of E_g. This is shown in the curves of Figure 5. The loss in R_h is usually much less than that in R_p due to the relative impedances of the two branches and the relative resistance of the two branches. (The plate resistance branch and the inductive leg.) As long as the tube is not capable of supporting self oscillations, the sum of these losses is greater than the power input from the tickler.

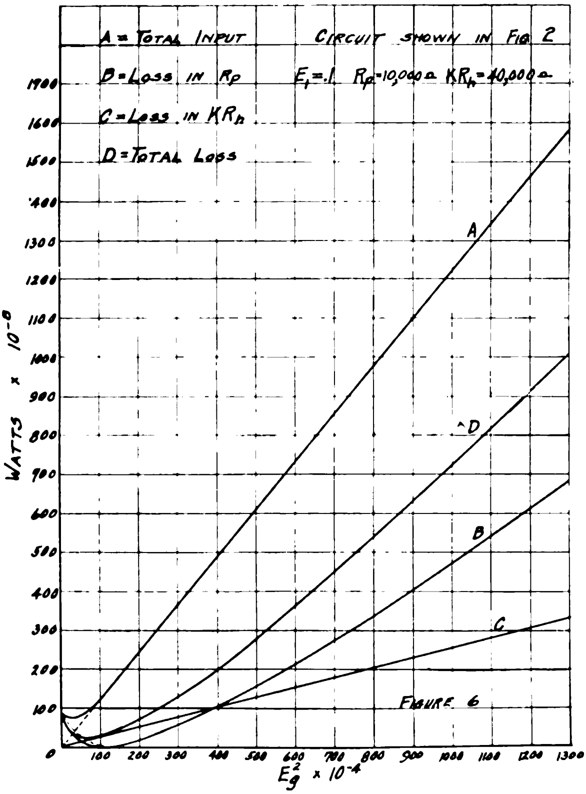


FIGURE 6

Now let E_1 be applied. At the instant of application E_1 supplies power at the rate shown in Figure 4. However, as E_g increases, the plate circuit also supplies power to the grid circuit as shown in Figure 5. The sum of these "power input" curves gives the total power input for any resulting value of E_g. This sum is shown in Figure 6. The loss in R_h is directly proportioned to E_g^2 and consequently this loss curve in Figure 6 is not different

than that shown in Figure 5. However, the loss in R_p is no longer the same as before. As E_g increases, the loss in R_p decreases, due, as previously shown, to the decreased voltage across R_p. When E_g equals E_1, the loss in R_p is zero. Now as E_g further increases, the loss in R_p increases, due to a greater increase of voltage across R_p. This loss curve is shown in Figure 6. The sum of the two loss curves is shown and it may be seen that (for the range of values shown) the input is greater than the corresponding losses and the tube will "oscillate." In oscillating, the grid voltage will build up until some point of power balance is again reached, i.e., where the losses increase to equal the input As soon as E_1 is removed, the additional loss incurred in R_p wil. bring back the conditions of Figure 5, where no oscillation will be produced. Regeneration therefore simply means one thing and one only. *When a certain voltage E_1 at a frequency f_1 is introduced into a circuit tuned to f_1, as shown in Figure 2, the losses occurring in the circuit will be sufficiently lowered to allow the electron tube to oscillate at a frequency f_1.* In so oscillating, it builds up and maintains a stable value of voltage E_g on the grid. *Regeneration and oscillation are one and the same thing.*

The equation for critical regeneration is as follows:

$$\frac{E_g^2}{R_p} + \frac{E_g^2}{K\,R_h} + \frac{E_g^2}{R_g} = \frac{E_t^2}{R_t} = \frac{b\,E_g^2}{R_t} \tag{1}$$

In this equation all terms are as previously stated except R_g, E_p, E_t, b, and R_t. R_g is the grid-filament resistance of the tube used as a regenerator, E_t is the voltage across the tickler coil, b is a constant of such a value that $\frac{b\,E_g^2}{R_t}$ expresses the value of tickler input to the grid and R_t is the effective resistance of the tickler coil measured across its terminals and is due to the absorption of power in the grid circuit. When E_1 is applied, the power furnished to the first tube by E_g is

$$e_g\,I_p = E_g\,\frac{(E_g - E_1)^2}{R_p} = E_g^2 - \frac{E_g\,E_1}{R_p} \tag{2}$$

As E_t is directly proportional to E_g (variations will be discussed later), all other terms remain constant. Subtracting (2) from the first term of (1) gives

$$\text{Watts Saved} = \frac{E_1\,E_g}{R_p} \tag{3}$$

Knowing E_1 and R_p, the total loss curve in Figure 6 with E_1 applied can be found by subtracting the above expression (3)

for "watts saved" from the total loss curve of Figure 5. If the tickler feed-back is that of critical regeneration and is a straight line function of E_g^2, i.e., where (1) holds true for all values of E_g, the loss curve with E_1 applied will always be less than the tickler input and the voltage E_g will build to an infinitely large value.

Reasons for Limited Amplification

Our next and perhaps most important step is to determine what limits the amplification of a regenerative electron tube, and how these limits can be controlled in practice. It may be seen from Figure 6, that there is no power input from E_1 after E_g equals E_1. This condition will be readily reached in practice, o for the present we will limit ourselves to a discussion of the imits with respect to E_g.

In a practical example of the circuit, R_h does not vary. However, the values of the plate impedance of both tubes and the input impedance of the regenerator change with changes in amplitude of E_g. The initial values of these impedances are greatly affected by the type of tube, direct current plate and grid voltages, etc. In addition, the amount of change as E_g varies, is greatly affected by the plate and grid voltage used. Another variable introduced is that due to the fact that the tickler input does not increase as a straight line function of E_g^2.

We have seen that when E_1 is introduced, the power input exceeds the power lost. The grid voltage E_g will build up to the point where the input and the loss are again equal. This means a loss increasing faster than E_g^2, or a tickler input increasing more slowly than E_g^2, or both.

These variations from straight line characteristics serve to bring about this new point of balance. The following equation expresses the power relation at the point of balance.

$$\frac{E_g(E_g-E_1)}{R_p}+f_p(E_g-E_1)+\frac{E_g^2}{K\,R_h}+\frac{E_g^2}{R_g}+f_g(E_g) = \frac{bE_g^2}{R_t}-f_{bt}(E_g). \qquad (4)$$

where $f_p\,(E_g-E_1)$ is a variable function of (E_g-E_1) expressing the difference between the actual power loss in the plate impedance with a finite E_g and the power loss as given by $\frac{(E_g-E_1)^2}{R_p}$: where $f_g\ (E_g)$ is a variable function of Ef, indicating a change in loss from that expressed by $\frac{(E_g^2)}{(R_g)}$: andwhere $f_{bt}\,(E_g)$ is a variable func-

tion of E_g, which includes all changes of the power supply from the tickler to the grid not in direct proportion to E_g^2.

Then

$$f_p(E_g-E_1)=\frac{E_g(E_g-E_1)}{R_p'}-\frac{(E_g-E_1)}{R_p} \tag{5}$$

$$f_g(E_g)=\frac{E_g^2}{R_p'}-\frac{E_g^2}{R_p} \tag{6}$$

$$f_{bt}(E_g)=\frac{b\,E_g^2}{R_t}-\frac{b\,E_g^2}{R_t'} \tag{7}$$

where R_p, R_g, and R_t are of values of the corresponding impedances when E_g is infinitesimal and where R_p', R_g', and R_t' are the values of impedance for finite values of E_g.

Notice that all of these functions are zero for $E_g=$zero, and can be positive or negative, depending upon the values of the primed numbers.

It is interesting to note that equation (4) holds good whether critical regeneration is used or not. This makes it very easy to check experimentally. A board set up of the circuit of Figure 2 was used with electron tube voltmeters to measure E_1 and E_g. With a given tickler adjustment E_1 and E_g were measured. Then E_1 was removed and the value of R_p increased by a value Y until E_g rose to its previous value due to self oscillation.

$$\frac{E_1\,E_g}{R_p} \text{ was then calculated,}$$

$$\text{and } \frac{E_g^2}{R_p}-\frac{E_g^2}{R_p+Y} \text{ was calculated.}$$

The two were found equal.

An interesting case is that at which the tickler is adjusted until $E_g=E_1$. The voltage across R_p is then zero and R_p may be varied without affecting E_g. When R_p is open circuited, the tuned circuit will be found to be oscillating with the same voltage E_g on the grid. This, too, has been experimentally verified.

If we assume critical regeneration we may subtract equation (4) from (1).

$$\frac{E_1\,E_g}{R_p}=f_p(E_g-E_1)+f_g(E_g)+f_{bt}(E_g). \tag{8}$$

The left-hand member of this equation represents the "watts saved" by introduction of E_1. The right-hand member represents the change in losses which utilize this "surplus" power at the point of balance.

Let us examine the limit of regenerative amplification as these functions are varied, keeping in mind the following:

First, in order for the tube to be in a state capable of "regenerating" but never capable of producing "self sustained" oscillations, the total loss curve must be greater than the tickler input for all values of E_g (no E_1 applied).

Second, critical regeneration is obtained where the loss curve and the tickler power input curve are tangent at $E_g = 0$.

Third, the maximum amplification is obtained when the abscissa of the point of intersection of the loss curve and the power input curve is a maximum with a given value of E_1.

Case I. When each of the above functions are equal to zero, or if the sum is equal to zero, or if the right-hand side is less than the left-hand side for all values of E_g, then E_g will increase without limit regardless of the value of E_1. This is true because under the terms of Case 1, the power input is always greater than the loss and E_g will increase indefinitely. Such a condition, although impossible practically, is shown in Figure 6.

Case II. Where the sum of the functions is finite and positive, and increases faster than the left-hand member of (8); then E_g will increase to a finite limit. This value depends on the value of E_1, but is not necessarily in direct proportion. This is shown in Figure 7. Upon removing E_1, E_g will drop to zero and the conditions given in equation (1) will be resumed.

Case III. Where the sum of all functions is negative or equal to zero for small values of E_g and become positive for larger values of E_g. In this case an infinitesimal value of E_1 will be sufficient to upset the balance of power and E_g will build up until the conditions of equation (8) are again obtained. This will occur when the sum of the functions reaches a sufficient positive value. This case is shown in Figure 8. Upon removing E_1 the voltage E_g will not fall to zero, but will remain at the point P', where the sum of the functions equals zero. To get the condition of critical regeneration again, it is necessary to reduce the tickler coupling and readjust it. This condition is known as "snapping" or "rubber-band effect" or "floppiness."

Limiting Effects Without Grid Leak and Condenser

The variations of the above functions with respect to E_g can be shown very clearly by means of curves. Figure 9 shows average curves of the variation of R_p with the variations of E_g. This grid voltage is superimposed on the direct current plate voltage of the amplifier tube.

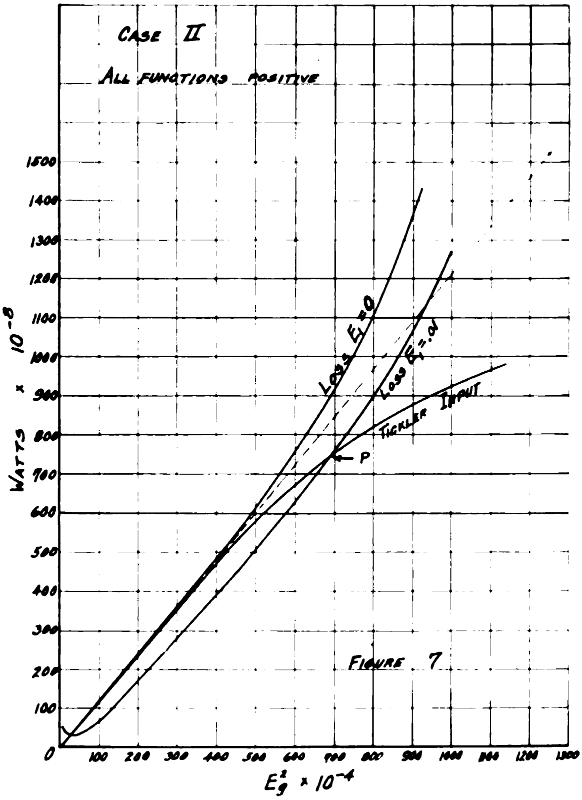


FIGURE 7

Figure 10 shows average curves of grid resistance with variations of amplitude of E_g. Various loss curves are plotted in Figure 11. These actual losses in watts are plotted against $E_g{}^2$, to produce as near a straight line curve as possible. However, in order to show the variations in loss curves by changing the circuit constants, extreme values have been chosen and the graphs are necessarily curved.

Figure 12 shows the variation in plate current with grid voltage. This is a direct current dynamic curve with resistance load in the plate circuit. From this curve are obtained the curves on Figure 13, which are alternating current dynamic curves. These curves indicate the alternating current plate current with alternating current grid volts. Notice that the grid bias very greatly affects the shape of these curves.

The power input to the grid circuit from the tickler is equal to I^2 (alternating current)$\times R_t$, where R_t is the previously men-

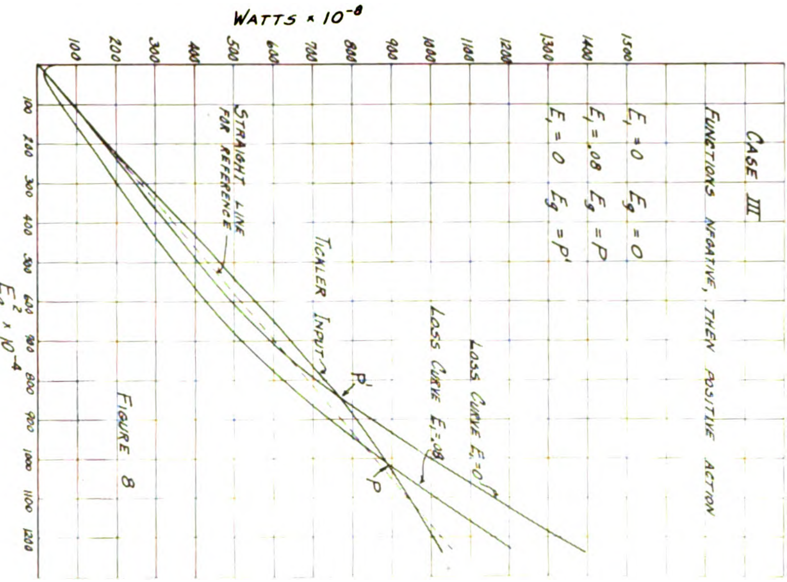


FIGURE 8.

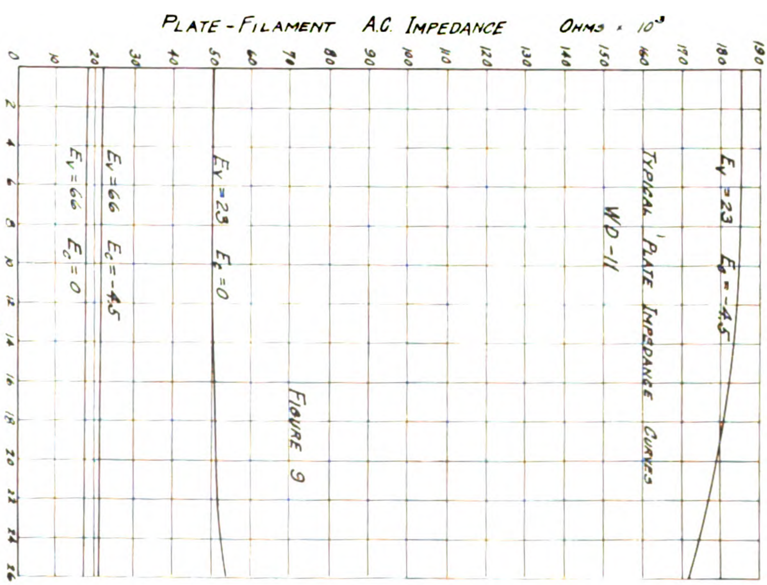


FIGURE 9

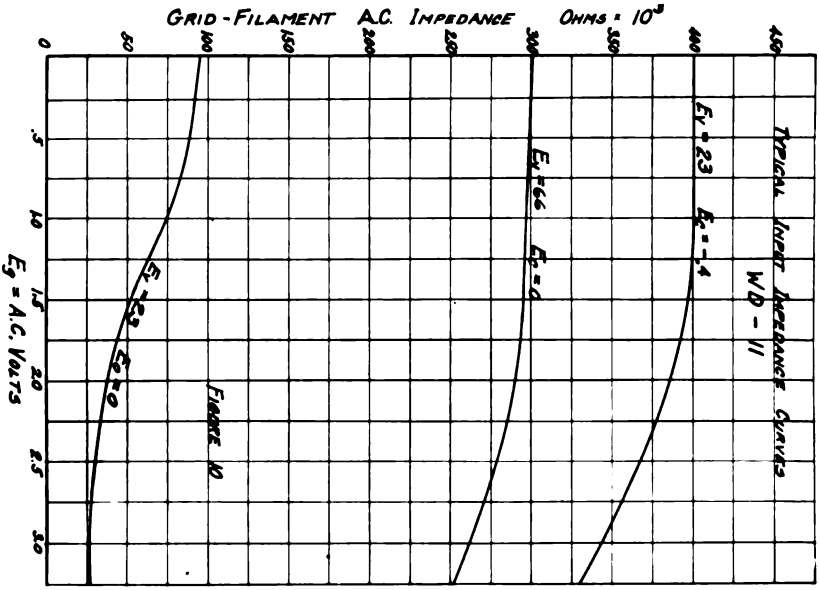


FIGURE 10

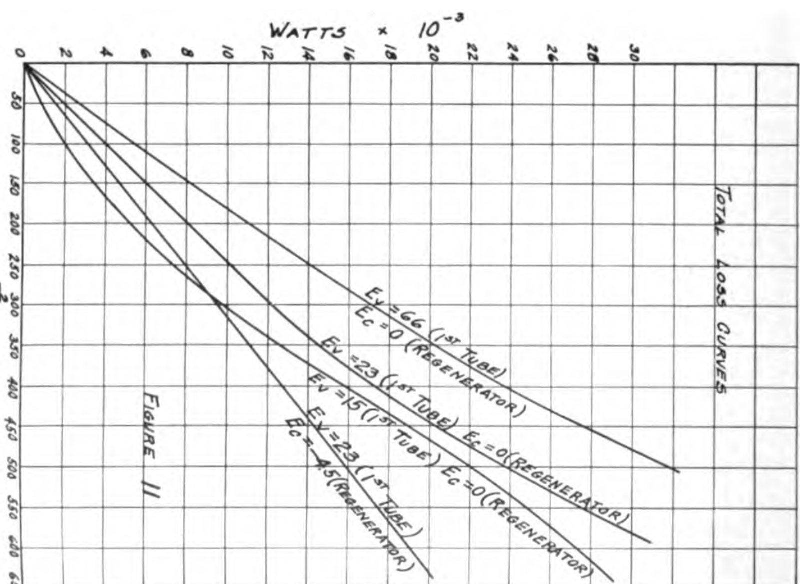


FIGURE 11

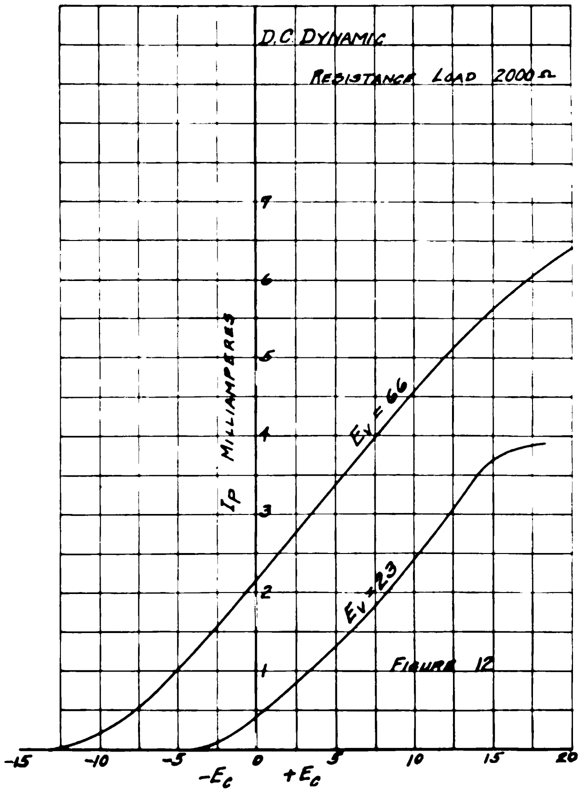


FIGURE 12

tioned resistance of the tickler coil, due to its coupling to the tuned circuit. The value of R_t increases directly as the tickler coupling is increased. I_p decreases slightly as R_t is increased, but as the value of R_t is low compared to the total impedance of the plate circuit, I_p does not change a large amount. Figure 14 shows several power dynamic curves. $E_g{}^2$ is used as abscissa to conform with the abscissa used on the loss curves; also to maintain the same shape of dynamic curves, (due to ordinates, $I_p{}^2 \times R_t$).

From these loss curves and input curves it is readily seen that practical conditions may vary greatly, and the limit of amplification varies with every change in operating conditions.

LIMITING EFFECTS WITH GRID LEAK AND CONDENSERS

In the regenerative action considered so far, the tube was used simply as an amplifier and not as a detector. It is common

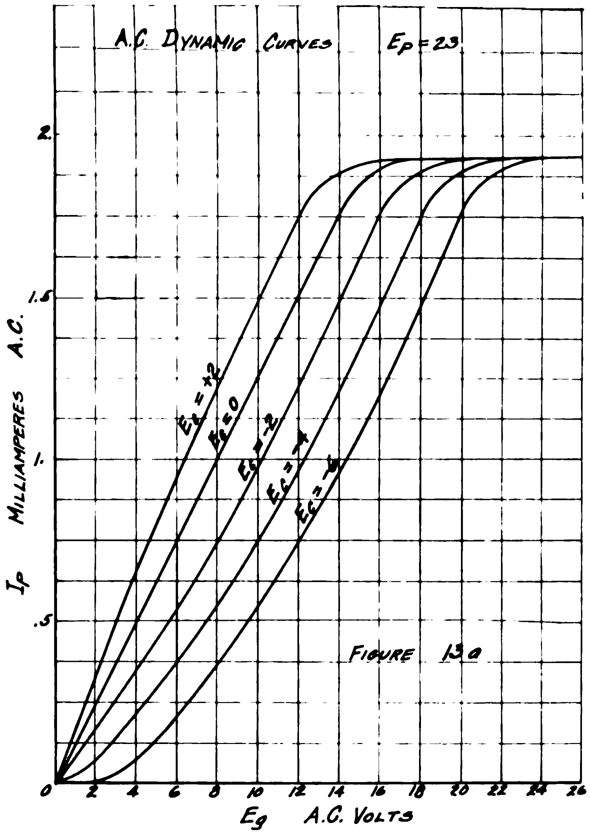


FIGURE 13a

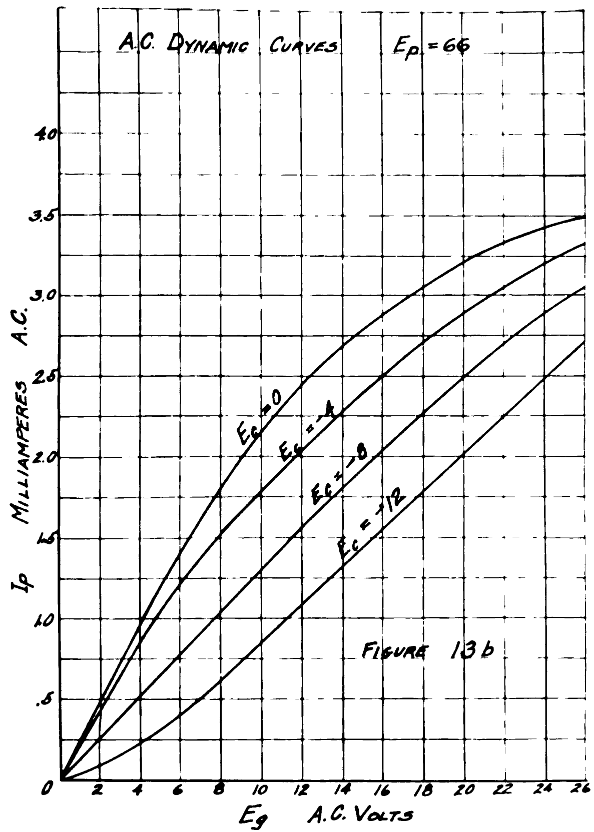


FIGURE 13b

practice, however, to regenerate the detector when using a grid leak and condenser. The component effects in this case are somewhat different than previously indicated.

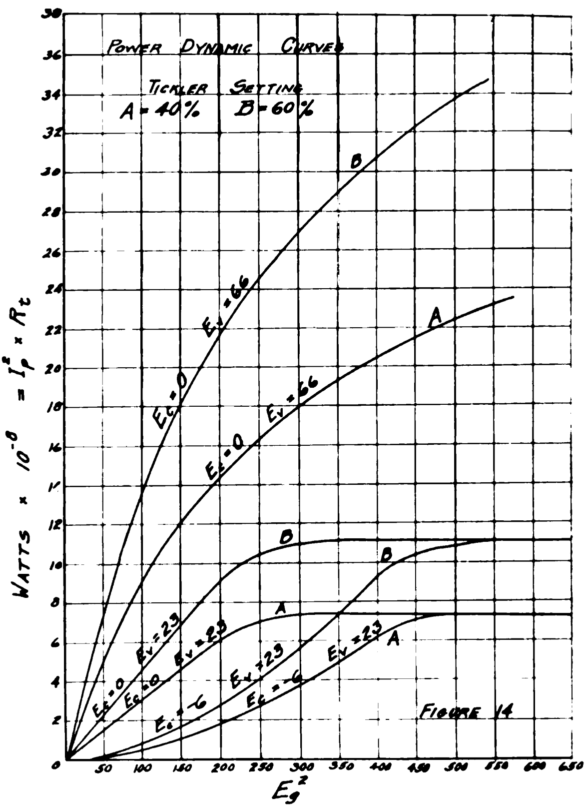


FIGURE 14

The first change is in the effective resistance of the grid leak and condenser plus the grid filament resistance. If the grid leak is small, the condition is not greatly different from no grid leak at all. That is, the current increases faster than E_{g} and the resistance loss increases faster than in direct proportion to E_{g}^2 If the grid leak is high, the grid condenser accumulates a negative change, tho the grid current must always increase as E_{g} increases, it will not increase as swiftly as in direct proportion. In this manner the effective resistance of the grid filament circuit increases and the loss does not increase as fast as E_{g}^2. These loss curves in the grid filament impedance, plotted against E_{g}^2, are shown in Figure 15, using two values of grid leak resistance.

The alternating current dynamic characteristic of power input is also different with grid leak and condenser. As E_g is applied, the average grid potential moves down on the static curve, thus increasing the impedance of the tube. Such a dynamic curve for 22½ volts (E_r) with grid leak and condenser is shown in direct comparison with a similar curve without grid leak and condenser in Figure 16. The reason for this maximum power

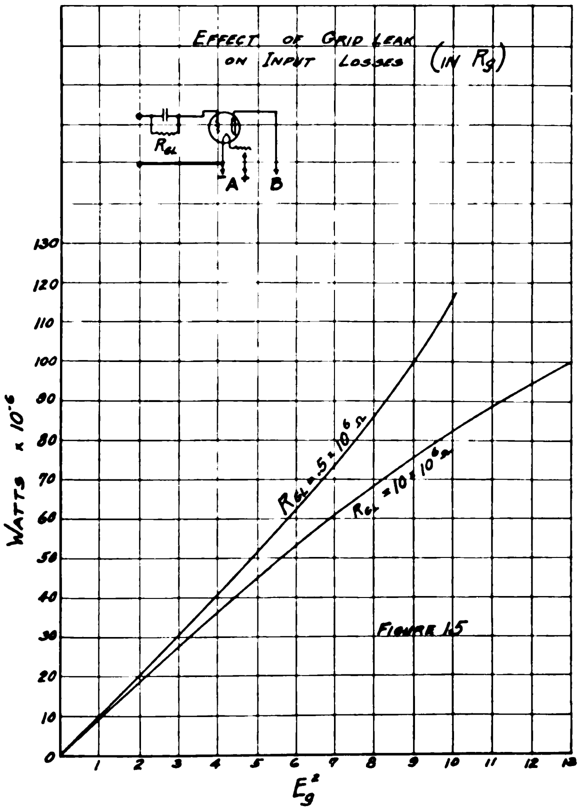


FIGURE 15

input change is seen in the static curve of Figure 17. When the grid draws current on the positive half cycle of E_g, the grid accumulates the negative charge and the average charge becomes increasingly negative. The grid becoming negative increases the tube impedance and so decreases the plate current.

With large values of plate voltage the change in plate impedance is less with a given change in grid bias. The result in the dynamic curve with grid leak and condenser is shown in Figure

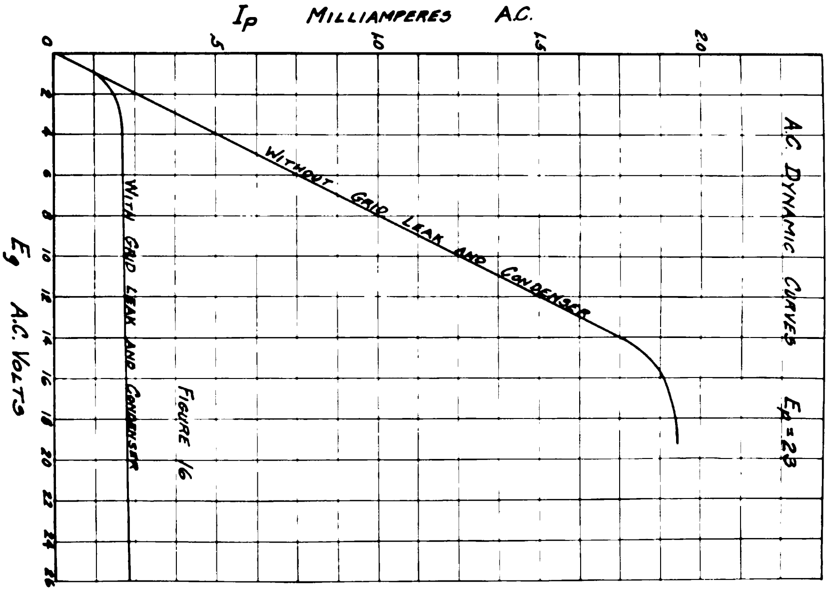


FIGURE 16

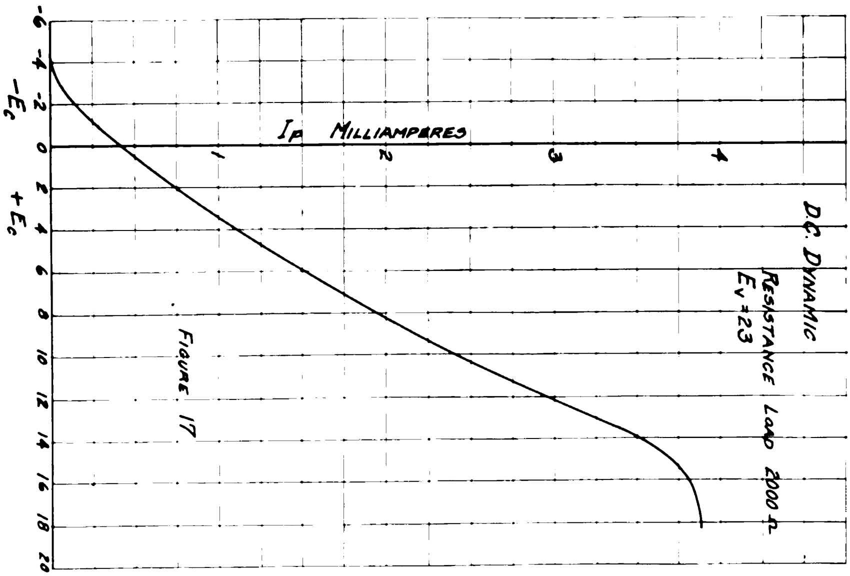


FIGURE 17

18. Comparing the static curve of Figure 19 with that of Figure 17 shows the reason for the relative changes in impedance.

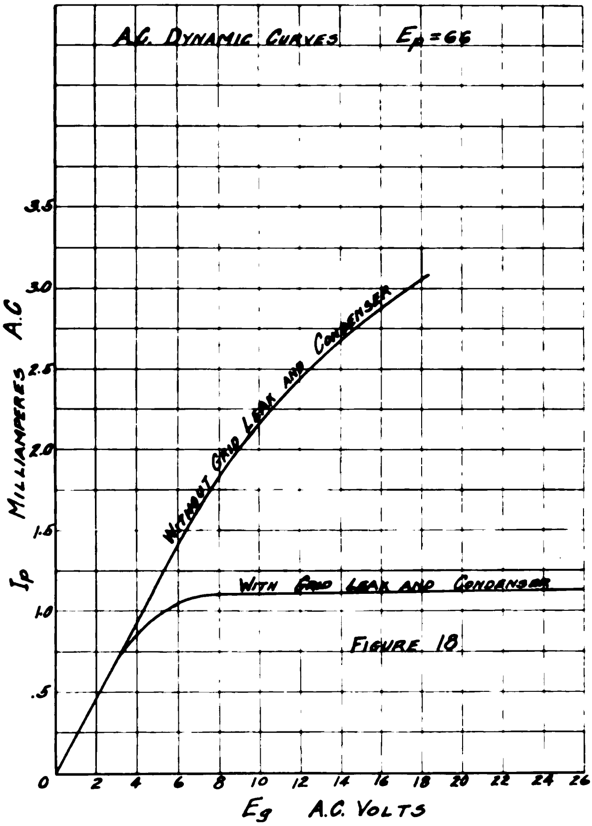


FIGURE 18

As previously pointed out, the maximum amplification will be obtained when the loss and input curves follow each other for the greatest distance. The dynamic curves for grid leak and condenser used under ordinary conditions (22½ volts on the plate), are always concave down. As seen from the loss curves we can adjust the grid leak to change the loss curve in R_g. If we continually adjust the tickler to maintain critical regeneration, decreasing the grid leak resistance increases the total loss curve, increases the tickler feed-back, and also decreases the rate of downward curvature of both curves. However, the curvature*

*In the following discussion the word curvature must be understood to mean curvature downwards. Thus if the curve under discussion is concave up, its curvature must be thought of as negative, and the phrase "decreasing its curvature" means increasing its rate of upward curvature.

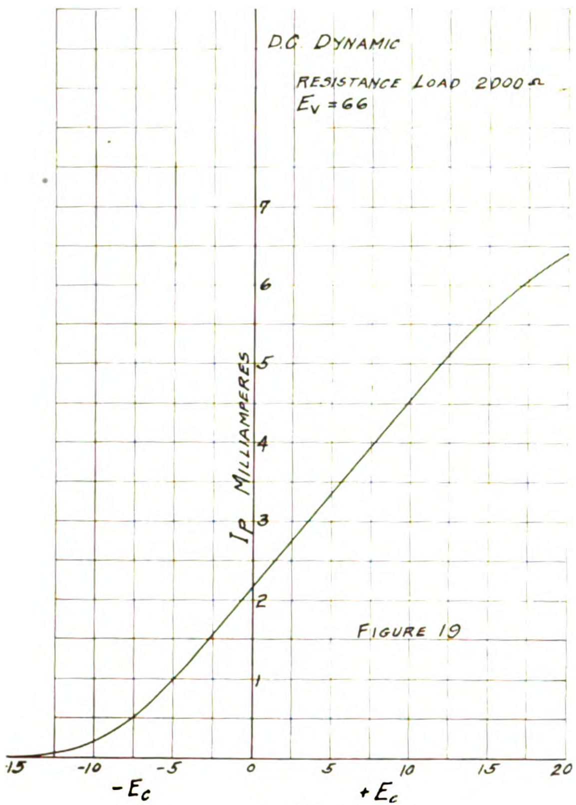


FIGURE 19

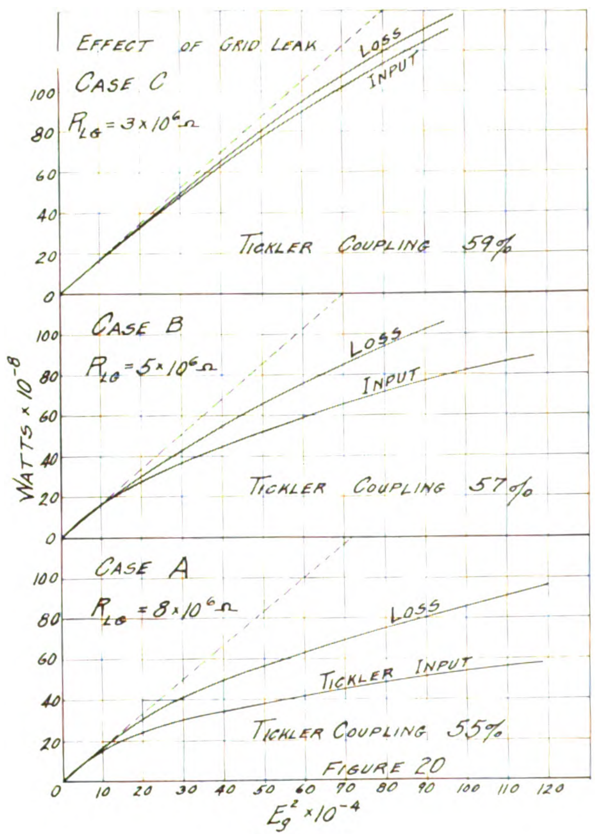


FIGURE 20

of the tickler input curve decreases faster than that of the loss curve. Thus, the two curves continually approach each other. This condition is shown in the curves of Figure 20.

It is readily conceivable that with certain values of grid leak, the curvature of the tickler input may be less than that of the total loss curve. Now if the point of critical regeneration is approached, that is, where the loss curve and input curve are tangent at zero, the tube will "snap" into oscillation. Such a condition is shown in Figure 21 and in Figure 22. The input curvature is less than that of the loss in the range shown. However, with large values of E_g the input curve becomes horizontal. The loss curve always increases so that at some point these two will be equal. The tube (in Case C, both Figures) will oscillate at this value of E_g, not shown on the curve.

The relation between losses and input with high grid leak

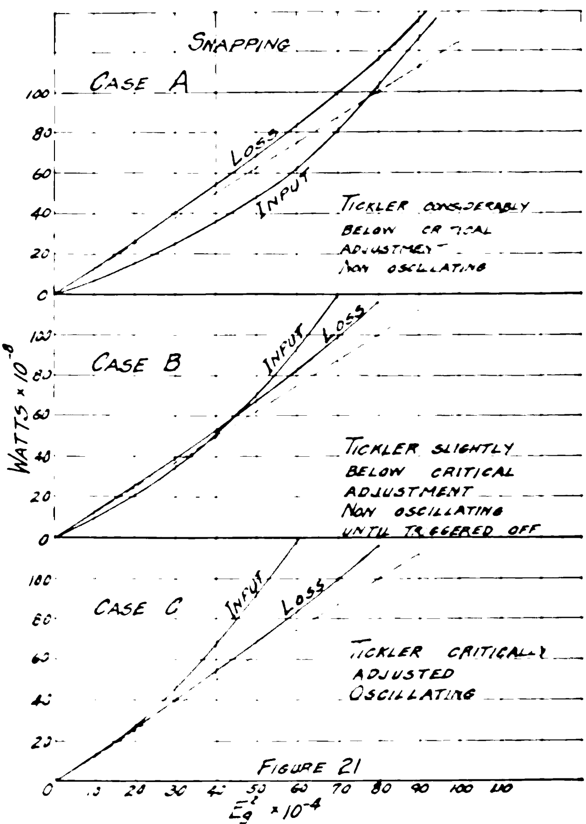


FIGURE 21

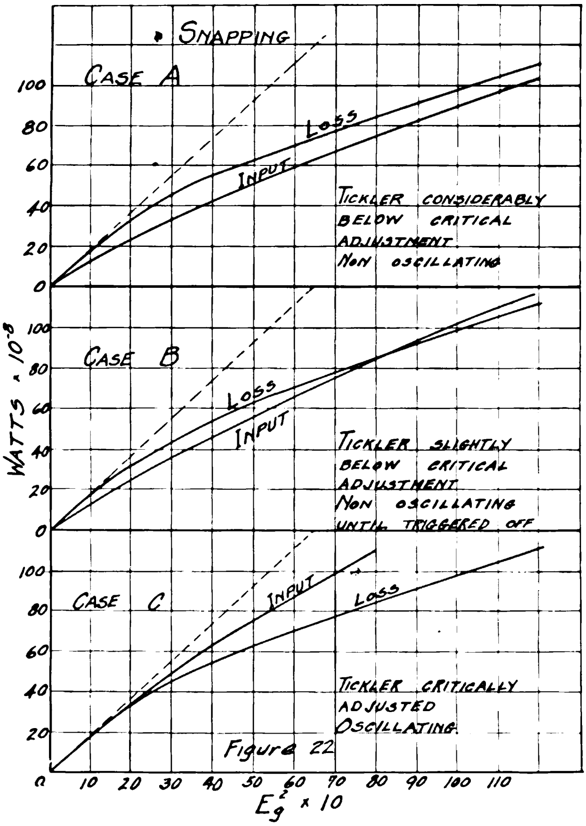


FIGURE 22

resistance shows that the two curves do not follow each other for a great distance and, therefore, the resultant amplification is low. Satisfactory operation cannot be obtained with snapping, as shown in Figure 22, so a low leak resistance will not produce any better results than a high resistance. Obviously there is some best value of grid leak to use with a regenerative detector. The results obtained when using this correct value explain in a large measure the growing use of a variable grid leak in sets where maximum sensitivity is desired on very weak signals, regardless of a multiplicity of adjustments.

AMPLIFICATION WITH AND WITHOUT GRID AND CONDENSER

Figure 23 shows how the form of dynamic curve with grid leak and condenser effects the voltage amplification. The curves are a self-evident demonstration of the higher amplification with-

out grid leak and condenser. These curves were taken at 22½ volts on the plate of the regenerator.

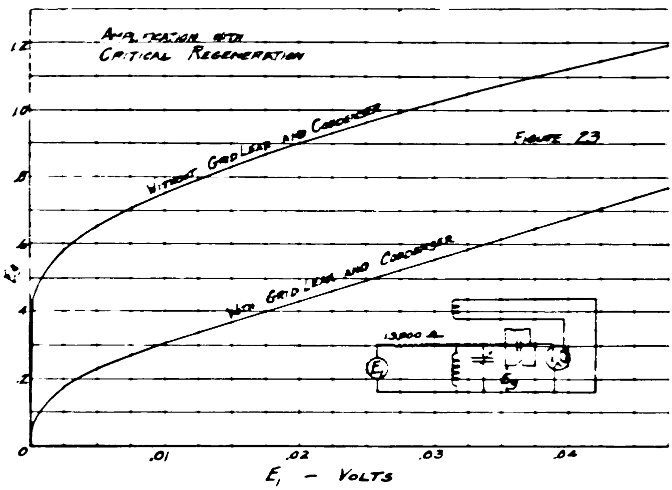


Figure 23

The decreased voltage amplification on the grid due to the introduction of the grid leak and condenser may or may not be compensated for by the increased detecting efficiency of the system. Thus the audibility of the signal as measured by an audibility meter in the plate circuit might be more with grid leak and condenser even tho a lower maximum voltage is obtained on the grid.

Effect of Resistance in the Grid Circuit

An important factor in the voltage obtained by regenerative amplification is the magnitude of the resistance in the grid circuit. So long as the tube can be made to oscillate, it is popularly supposed that the same voltage may be obtained at critical regeneration, regardless of the grid circuit resistance. Let us examine the actual results obtained by changing the resistance.

Assume the circuit of Figure 2 to be adjusted to critical regeneration and that E_1 is applied. Equation (4) will then express the power balance,

$$\frac{E_g(E_g-E_1)^2}{R_p}+f_p(E_g-E_1)+\frac{E_g^2}{K\,R_h}+\frac{E_g^2}{R_g}+f_g(E_g) = \frac{b\,E_g^2}{R_t}-f_{bt}(E_g) \qquad (4)$$

and a stable value of E_g will be obtained.

Now assume that E_1 is removed, R_h is increased and the tickler readjusted to critical regeneration. Assume that E_1 is now reapplied and also that the same value of E_g is momentarily existant. If the power loss under these conditions is less than the power supplied by the tickler, the voltage will build higher until a new condition of balance is obtained. If the power loss is greater than that supplied by the tickler, then we have chosen too large a value, and E_g will be smaller in a condition of balance than we have assumed to be momentarily existant.

An examination of equation (4) shows that the only terms affected by the changed conditions are:

$$\frac{E_g^2}{K R_h}; \quad \frac{b E_g^2}{R_t}; \quad \text{and } f_{bt}(E_g)$$

But since the tickler has been readjusted for critical regeneration with the increased resistance, the change in $\frac{E_g^2}{K R_h}$ has been exactly counterbalanced by the change in $\frac{b E_g^2}{R_t}$. Hence the only term tending to upset the equality is $f_{bt}(E_g)$. With a given E_g if the tickler input represented by $\frac{b E_g^2}{R_t}$ is increased by increasing the tickler coupling, then $f_{bt}(E_g)$ will be increased.

If the function is positive (that is, if the tickler curve is concave downward as in Figure 24), then the power loss exceeds the power input under the second set of conditions and the true balance will be obtained with a smaller value of E_g than that given by the first set of conditions. *This means that increasing the value of R_h will decrease the value of E_g even tho c.itical regeneration is obtained.*

Conversely, if this function is negative, E_g will build to a larger value in the second case than in the first. This case would seldom occur in practice. Notice, however, that the function being negative means that the tickler curve is concave upward. To obtain critical regeneration without self oscillation, under such conditions, the loss curve must have a larger rate of curvature upward than the tickler input curve. *Under these conditions, an increase in R_h would produce greater voltage amplification.*

If the function is zero (that is, if the tickler input curve is a straight line), the value of R_h will have no effect. This condition may be approximated in practice. In most cases, however, an

increase in circuit resistance decreases the available amplification.

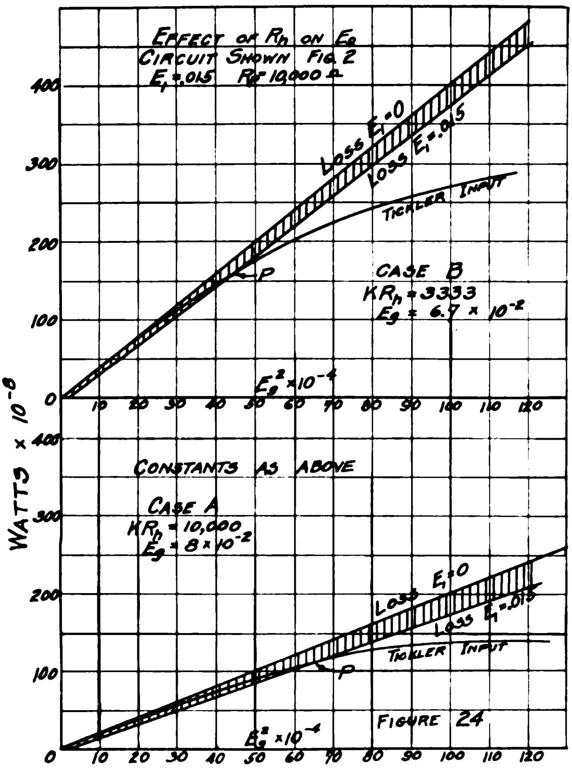


FIGURE 24

An illustration with curves will make the point clearer. Assume the circuit to be so adjusted that the loss curves are straight lines and the dynamic curve increases as shown in Case A, of Figure 24. Let the resistance loss be doubled by increasing R_h, as shown in Case B. Now double the value of tickler coupling. The effective resistance R_t is doubled, but the power input is not, due to a decrease in I_p, with increased plate circuit impedance. Thus in order to obtain the point of critical regeneration with the increased resistance losses, the tickler coupling must be more than doubled.

Let a certain value of E_1 be applied to the low loss circuit, in Case A. The new total loss curve is as shown, the area between the loss curves being slightly shaded. The lower loss curve can be obtained for any value of E_g with the given value

of E_1, by subtracting the value given by (3) from the upper loss curve. The point of power balance, or the final value of E_g, is indicated.

Now consider the result when the same E_1 is applied to the higher loss circuit. The new loss curve lies the same distance below this upper loss curve as it did in the previous case. The value of "watts saved" depends only on E_1 and E_g, and is independent of R_h. This new loss curve will intersect the input curve at a very much lower value of E_g. Thus in this case increasing the resistance will decrease the voltage amplification (E_g/E_1), even tho the point of critical regeneration is obtained for the larger resistance.

Dynamic curves such as those shown in Figure 24, are obtained under normal conditions when using a grid leak and condenser. This indicates that under operating conditions, the resistance may very greatly affect the voltage amplification. Figure 25 shows the circuit which was set up to test this statement and the curve resulting. It can be readily seen that the grid voltage at critical regeneration decreases as the resistance losses in the oscillatory circuit are increased.

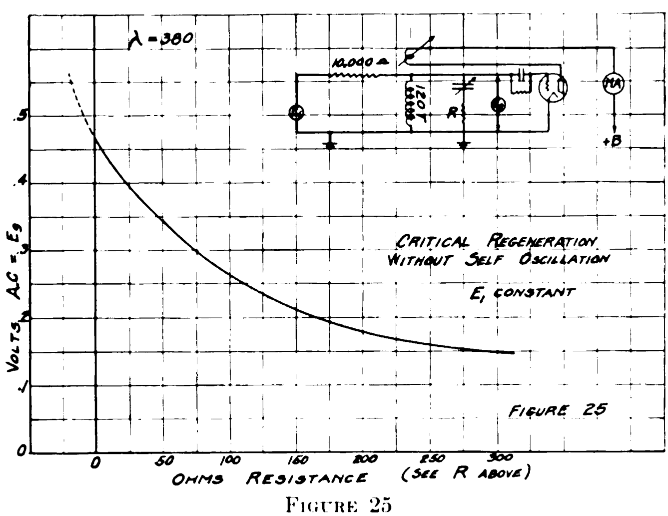


FIGURE 25

TURN RATIO FOR MAXIMUM AMPLIFICATION

Regenerated transformers often use a step-up ratio. As in the case of unregenerated transformers, there is a certain turn ratio which will give maximum amplification.

Consider the action in the circuit of Figure 2 if a step-up ratio

is used. The equation for a power balance with critical regeneration and E_1 applied is

$$\frac{N E_g (N E_g - E_1)}{R_p} + f_p(N E_g - E_1) + \frac{E_g^2}{K R_k} + \frac{E_g^2}{R_g} + f_g(E_g) = \frac{b E_g^2}{R_t} - f_{bt}(E_g). \tag{9}$$

where $N E_g$ is the voltage that would appear across the primary coil if the circuit were oscillating with E_g volts on the grid. Then N is the effective turn ratio of the primary to the secondary of the transformer.

$$\text{Watts saved } \frac{N E_g E}{R_p} = f_p(N E_g - E_1) + f_g(E_g) + f_{bt}(E_g).$$

For simplicity let us assume $f_p(N E_g - E_1) + f_g(E_g)$ is zero. In other words, the plate impedance of the first tube and the grid input impedance of the second tube do not vary as E_g changes. The effect of these on the result will be discussed later

$$\text{Then } f_{bt}(E_g) = \text{watts saved} = \frac{N E_g E_1}{R_p} \tag{10}$$

If we subtract "watts saved" from both sides of equation (9) and combine the terms

$$\frac{E_g^2}{K R_k} \text{ and } \frac{E_g^2}{R_g} \text{ into } \frac{E_g^2}{Z}$$

(where Z is the impedance of the tuned circuit across the terminals of the condenser when R_p is removed), we then have

$$\frac{N^2 E_g^2}{R_p} + \frac{E_g^2}{Z} = \frac{b E_g^2}{R_t} \tag{11}$$

If we vary the ratio N and continuously readjust for critical regeneration, the rate of change of E_g with respect to N is zero at the point of maximum amplification. If we derive the above equation, letting $\frac{d E_g}{d N} = 0$, and solve for N, we get a formula for the ratio giving maximum amplification. We must keep in mind, however, that the rate of change of $\frac{b E_g^2}{R_t}$ is not zero at this point, since the tickler is continuously readjusted. If R_t is small, compared to the tube impedance, $\frac{b E_g^2}{R_t}$ is directly proportional to $f_{bt}(E_g)$, hence their derivatives are proportional in the same ratio. That is

$$\frac{\frac{d}{dN}\frac{b\,E_g^2}{R_t}}{\frac{d}{dN}f_{bt}(E_g)} = \frac{\frac{b\,E_g^2}{R_t}}{f_{bt}(E_g)} \tag{12}$$

$$\frac{d}{dN}\frac{bE_g^2}{R_t} = \frac{b\,E_g}{R_t}\frac{\frac{d}{dN}f_{bt}(E_g)}{f_{bt}(E_g)} \tag{13}$$

The derivation of (10) gives

$$\frac{d}{dN}f_{bt}(E_g) = \frac{d}{dN}\frac{N\,E_g\,E_1}{R_p} = \frac{E_g\,E_1}{R_p} \tag{14}$$

Substituting in equation (13), the values given by equation (11), by equation (14) and equation (10), gives

$$\begin{aligned}\frac{d}{dN}\frac{b\,E_g^2}{R_t} &= \left[\frac{N^2\,E_g^2}{R_p} + \frac{E_g^2}{Z}\right]\frac{E_1\,E_g}{R_p}\,\frac{R_p}{N\,E_g\,E_1}\\ &= \left[\frac{N^2 E_g^3}{R_p} + \frac{E_g^3}{Z}\right]\frac{1}{N}\end{aligned} \tag{15}$$

Hence deriving equation (11), we get

$$\frac{2\,N\,E_g^2}{R_p} = \left[\frac{N^2\,E_g^3}{R_p} + \frac{E_g^2}{Z}\right]\frac{1}{N} \tag{16}$$

Simplifying (16).

$$\frac{2\,N^2\,E_g^2}{R_p} = \frac{N^2\,E_g^2}{R_p} + \frac{E_g^2}{Z} \tag{17}$$

Solving (17) for N, gives

$$\frac{N^2}{R_p} = \frac{1}{2} \tag{18}$$

$$N = \sqrt{\frac{R_p}{Z}} \tag{19}$$

Hence the best ratio is approximately $\sqrt{\frac{R_p}{Z}}$. It can be proved that this ratio is one which will give a primary impedance equal to R_p. This is the same ratio as that which would be chosen for an unregenerated transformer.

It was assumed in the above that $f_p\,(N\,E_g - E_1)$ and $f_g\,(E_g)$ were equal to zero. If $f_g\,(E_g)$ has a positive value, however, the result is changed toward a one-to-one ratio. $\frac{d}{d\,N} f_M(E_g)$ will not be affected since $\frac{d}{d\,N} f_g(E_g)$ is zero at the point of maximum amplification.

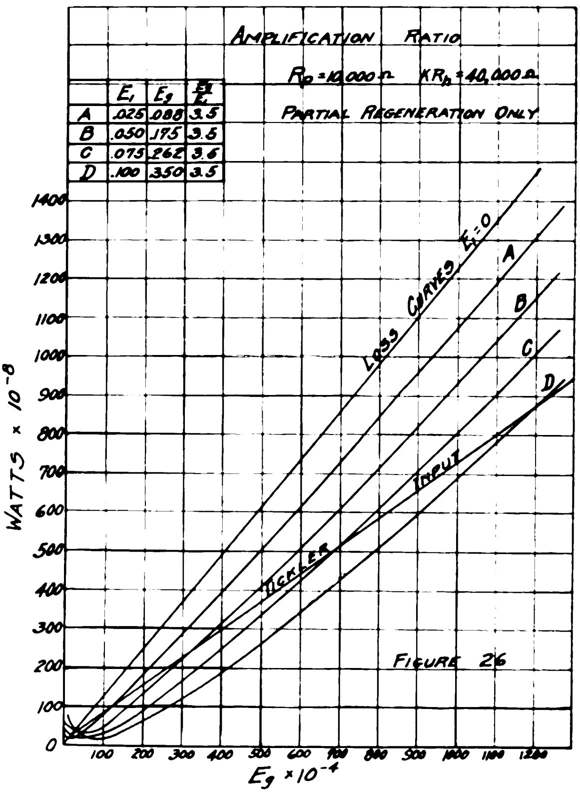


FIGURE 26

Normally $f_p\,(N\,E_g - E_1)$ will be so small as to affect the result only slightly.

It is evident from equation (19) and the above discussion, that for maximum amplification in a regenerated transformer, the correct ratio between secondary and primary turns $\left(\text{that is, } \frac{1}{N}\right)$ is equal to the best ratio of the transformer when unregenerated.

Variation in Amplification with Varying Input

A subject of much discussion concerning regeneration and one which muct be included here, is that of the ratio between the initial and the final signal, that is, the ratio between the grid voltage on the regenerating tube and the voltage on the grid of the radio frequency amplifier.

Two conditions are possible in this consideration, first, when the tube is capable of regenerating, and second, when it is adjusted to critical regeneration. Let us consider the action when the tube is capable of regenerating but not adjusted critically. The loss curve and the tickler input curve are assumed to be straight lines. That is, all functional terms in equation (4) are zero. If we let

$$\left(\frac{1}{R_p}+\frac{1}{K R_h}+\frac{1}{R_g}-\frac{b}{R_t}\right)=+S$$

the solution for E_g is

$$E_g=E_1\frac{1}{S R_p} \tag{20}$$

This means that E_g is *directly proportional* to E_1 for all values of E_1. Or in other words, the amplification ratio is independent of the strength of the applied signal. This is one of the conclusions reached in the "Bureau of Standards Sci. Papers," Number 487. Figure 26 shows loss curves and the corresponding values of E_g with a given tickler adjustment. Note that the amplification ratio is a constant.

However, there is one fallacy in this proof. The tickler input curve is seldom a straight line, as was assumed. It is usually concave downward. It is evident that under these conditions doubling the value of E_1 will not double the value of E_g. The actual ratio for any value of E_1 will depend on f_{bt} (E_g), or on the difference between the actual input curve and a straight line. Thus under normal conditions a greater amplification ratio will be obtained when using a small value of E_1. A close inspection of the observed points in Figure 3 of the above mentioned "Bureau Sic. Paper" *shows this actually to be the case in their results.* This apparent discrepancy was evidently attributed to experimental error and no attempt made at explanation.

While this variation from a constant amplification ratio is almost immeasureably small when low values of "feed back" are used, this effect becomes of supreme importance when the circuit is adjusted to critical regeneration. Such a condition is

shown in Figure 27. The sharp bend in the tickler input is somewhat exaggerated, altho this can be approximated under some conditions. Due to the enormous amplification obtained at critical regeneration, smaller values of input voltage were used for these curves. The loss curves for various values of applied signal are shown and the resulting values of E_g are noted. In this case it may be seen that the amplification ratio is almost *inversely* proportional to the strength of applied signal. This conclusion was previously advanced by N. C. Little, in his article on "The Limit of Regeneration."

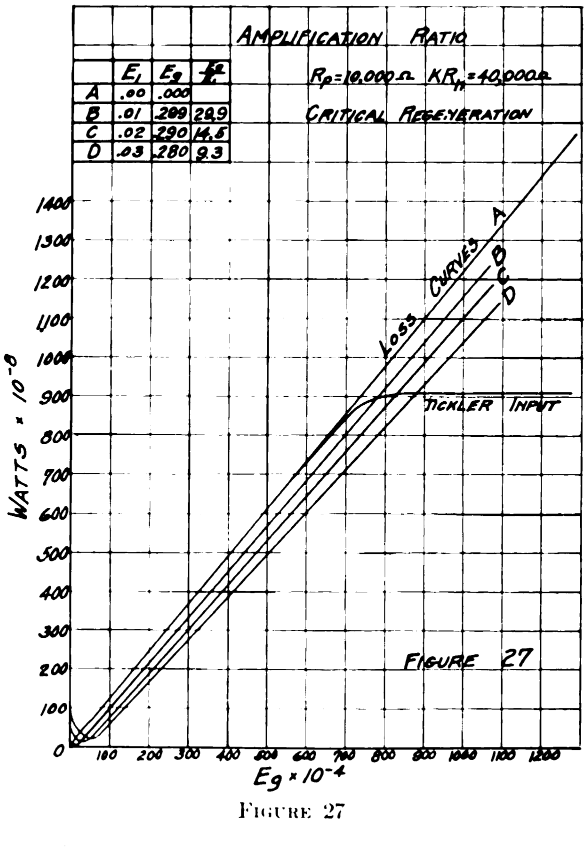


FIGURE 27

This effect accounts for the possibility of receiving distant stations when the amount of received energy is extremely small. The amplification ratio with such a weak signal may be as much as ten thousand or more at critical regeneration.

Figure 28 shows curves actually obtained by measurement

between amplification ratio and the strength of applied signal. Notice how rapidly the amplification ratio increases as the strength of applied signal decreases. The curves also show the difference in amplification ratio using a grid leak and condenser. This result could have been predicted by considering the change in the shape of the tickler input curve when using a grid leak and condenser.

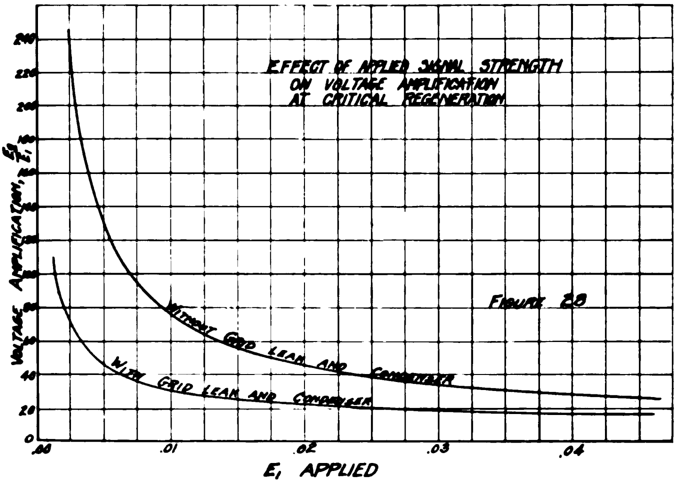


FIGURE 28

One result of this discussion is to show how greatly inferior radio frequency amplification is with respect to regeneration, especially when receiving distant and weak signals. With moderate signal strength a radio frequency amplifier may give a stronger output than a regenerative receiver, but if the signal input is reduced continuously a point will eventually be reached where the regenerative set will excel the radio frequency amplifier. The very nature of regeneration in amplifying weak signals more than strong ones is an effect which is more than we would honestly dare to require..

CIRCUIT CONDITIONS AS AFFECTING DISTORTION

In all of the above discussions we have assumed that the voltage E_1 was applied for an unlimited time and we have discussed the value of E_g after it had reached a stable condition. Under this consideration the L/C ratio (if $K\ R_h$ remains the same) will not affect the above results in the least. However, if the time of application is limited or if E_g must follow the amplitude

variations of a modulated continuous wave input, then the L/C ratio is important, since it affects the rate of increase of E_g.

We see from the expression (3) that with a given E_1 (where E_1 is small), the value of "watts saved" increased almost directly as E_g. We have also seen that as E_g increases, the losses increase until a point of balance is reached. Until this point is reached the watts saved are greater than the losses. In considering the time relations involved we can assume that the average available watts for building up the voltage E_g is a small part of the value given by (3). These average available watts are used to charge the condenser as E_g increases.

The bearing of the above on the effect of the L/C ratio can best be shown by concrete example, assuming plausible values for the circuit constants.

Let $E_1 = 0.1$ v. $E_g = 2.0$ v. at the point of balance.

Let $C = 100$ mmfd. $F = 1{,}000{,}000$ cycles per second.

$R_p = 10{,}000$ ohms.

$$\text{Then } \frac{E_1 E_g}{R_p} = \frac{0.29}{10{,}000} = 29 \times 10^{-6} \text{ watts}$$

This is the value of watts saved at the point of balance due to E_1. Then 10^{-5} is approximately the average value during the rise of E_g.

Assume that about 1/5 of the average watts saved are available for charging the condenser, the rest going to supply the extra losses due to the change in the functional terms as E_g increases. Then the average value of power available for charging the condenser will be about 2×10^{-6} watts. The energy stored in the oscillating circuit when $E_g = 2$ v. is given by the expression

$$1/2\, C E_g^2 = \frac{100. \times 2}{2 \times 10^{12}} = 2 \times 10^{-10} \text{ watt seconds.}$$

$\frac{2 \times 10^{-10}}{2 \times 10^{-6}}$ equals 1×10^{-4} seconds. This gives the time for the voltage to build up under the above assumptions. This time is equivalent to 100 cycles at the assumed frequency.

Of course it cannot be said that E_g builds to a final value in 100 cycles, since mathematically it never reaches the stable condition but approaches the final value asympototically. In other words, as the circuit approaches the stable condition, the rate of increase of E_g decreases so that it continually approaches, but never reaches its limiting value. In a small fraction of a second, however, E_g approximates its limiting value so closely that the difference would be too small to measure. This small fraction

of a second necessary for any value of E_1 to charge the condenser to the value determined by E_g, determines the maximum rate at which E_g can be changed and still obtain its maximum value. The inverse of this fraction gives the approximate frequency of modulation at which a decrease in amplification will become noticeable. In the case discussed this would be about 10^4 cycles. Higher frequencies would suffer greater decrease in amplification and lower frequencies scarcely at all. This explains why in general regeneration causes but little distortion.

If E_1 is decreased, E_g will decrease. If E_1 and E_g decrease in the same proportion, the watts available for charging the condenser will decrease in proportion to E_g^2. The energy in the condenser is directly proportional to E_g^2. Consequently the time required (and so the number of cycles) to charge the condenser is the same regardless of E_1. However, if E_g does not decrease as fast as E_1 (usually the case), the watts available for charging the condenser will be less in proportion. Thus to obtain the maximum voltage (approximately), longer time will be required when E_1 is small. In this case modulation frequencies below 10,000 cycles may suffer loss of amplification. This means that with critical regeneration, weak signals will suffer more distortion, due to the disproportionate amplification of the lower frequencies than will strong signals. This fact has often been observed in receiving voice modulated continuous waves.

If the circuit is tuned to another frequency by changing L, the energy in the condenser for given balance point (E_g) is constant. In this case the time required is independent of the carrier frequency and there is no more tendency to distort the modulation on low carrier frequencies than on high frequencies. If, however, the capacity is increased to tune to a lower frequency (higher wave length), more time in proportion will be required to obtain an approximate maximum E_g, and consequently the modulation will suffer greater distortion.

Optimum Heterodyne

It is generally understood that regenerative amplification only occurs when, independent of the signal, there are no local oscillations produced. It is nevertheless a fact that when the circuit is oscillating weakly, as in heterodyne reception, very great amplification is obtained of the beat between the signal and the local oscillations. This amplification is due to regenerative action.

Consider the action taking place when the circuit of Figure 2

is oscillating at a frequency f_2. The equation (4) previously given (using $E_1=0$), will give the equation of balance. Let E_1 be applied at a frequency f_1. At the instant of balance when E_1 and E_g are 180° out of phase, the decrease in power loss is given by

$$\frac{2\,E_1\,E_g}{R_p} \tag{3}$$

and as a result E_g will tend to increase as previously explained.

Now assume the point where E_1 and E_2 are in phase. The losses in R_p increase, due to the increased voltage across it. In this case E_1 is supplying power to the circuit.

The value of the loss in R_p when E_1 and E_g are in phase is given by

$$\text{watts in } R_p = \frac{E_g^2+2\,E_1\,E_g+E_1^2}{R_p}$$

However, E_1 is supplying part of this.

$$\text{Power supplied by } E_1 = \frac{E_1^2+E_1\,E_g}{R_p}$$

$$\text{Power supplied by } E_g = \frac{E_g^2+E_1\,E_g}{R_p}$$

Extra watts supplied by E_g, due to application of E_1 is $\dfrac{E_1E_g}{R_p}$ = Extra watts loss due to E_1.

It is evident that E_g will decrease and increase at the frequency (f_1-f_2), and the amplitude of this variation depends on the values of the "watts saved" and the "watts lost." It has previously been shown that for a given E_1 a low loss tuned circuit produces more amplification (a larger E_g). In addition, in heterodyne reception we want the "watts lost" to also be as large as possible. It is evident that the power available for changing E_g increases, if the amplitude of local oscillation is increased. The rate at which the power loss and tickler input curves diverge will also increase as the initial value of E_g is increased. This has a tendency to decrease the amplification.

As these two tendencies oppose, there is obviously an optimum initial value of E_g which will give the maximum response at the heterodyne frequency.

The efficiency of detection as affected by the initial value of E_g must also be considered in a complete analysis. The optimum point can only be found by trial.

The amplitude of the heterodyne beat will not be affected as f_1 and f_2 are separated, except as the time required to change

the charge on the condenser becomes appreciable. When the beat frequency is high, this effect noticeably decreases the amplitude of the beat.

"Locking-in" and Zero Beat Reception

In considering the effect of local oscillations with respect to regeneration, another well-known phenomena is to be considered. This is the effect of "locking in," where "locking in" refers to the dead spot observed in the center of the heterodyne beat note obtained between the local oscillator and the incoming signal.

Consider the circuit shown at "M," Figure 29. This is equivalent to Figure 2, except that L and C are tuned to a frequency f_2, and the tickler coupling has been increased until the tube is oscillating with the stable value E_g on the grid. E_1 is applied at frequency f_1.

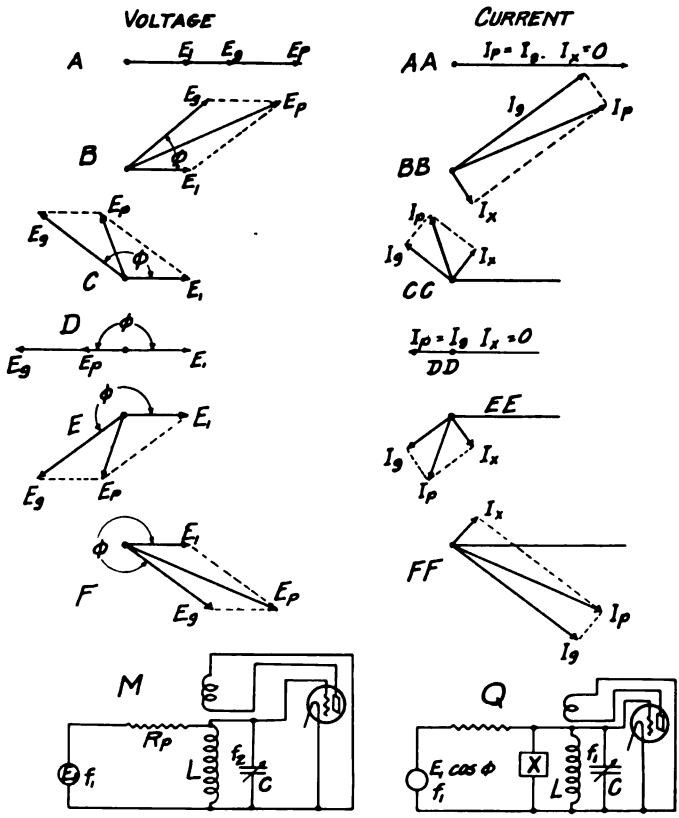


Figure 29

Let us assume an initial condition such as shown by the vectors at A. E_1 and E_g are in phase and the current in R_p is in phase with E_g (E_g is assumed slightly larger than E_1). Assume f_2 to be greater than f_1.

Then the phase relation between E_1 and E_g will change and the phase angle ϕ will increase. The phase relation for different values of ϕ are as shown at B, C, D, E, or F. The voltage E_p across the resistor and the current I_p thru it, are no longer in phase with E_g. As shown at BB, CC, DD, EE, and FF, however, the current has one component I_g, in phase with E_g and another I_x at right angles. Now

$$I_p^2 = \frac{E_g^2+E_1^2+2\,E_g\,E_1 \cos\phi}{R_p^2} = I_g^2+I_x^2$$

$$I_g = \frac{E_g+E_1 \cos\phi}{R_p}$$

$$I_x = \frac{E_1 \sin\phi}{R_p}$$

I_p is out of phase with E_g. Hence if an equivalent circuit is desired, in which the applied voltage is 180° out of phase with E_g, a reactance "X" must be shunted across R_p. Such an equivalent circuit is shown in Figure 29 at "L."

The value of "X" must be such that with E_g volts across it, a current of I_x amperes will flow thru it. Then

$$X = \frac{E_g}{I_x} = \frac{E_g\,R_p}{E_1 \sin\phi}$$

The current in R_p in this equivalent circuit is

$$I_g = \frac{E_g+E_1 \cos\phi}{R_p}$$

and the value of $I_p^2 = I_x^2 + I_g^2$ has not been changed. Hence X is an equivalent reactance introduced into the circuit by the phase relation between I_p and E_g.

Let us take a condition when f_2 is larger than f_1. Then if ϕ lies in the first or second quadrants, $\sin\phi$ is positive and X is positive (that is, an inductive reactance). Under these conditions a lagging current is flowing, as shown at BB and CC. The equivalent reactance X being inductive, the circuit is, in effect, tuned to a higher frequency and the value of f_2 is increased even farther above f_1. Hence the rate of change of ϕ is increased.

If ϕ lies in the third or fourth quadrants, then $\sin\phi$ is negative and X is negative (that is, a capacitive reactance). Hence f_2

is decreased in value toward f_1. This condition (that is, leading current) is shown at EE and FF.

If $f_2 - f_1$ is not too large before E_1 is applied, it is evident that after E_1 is applied, X will have some value X_o at which f_2 is changed so that $f_2 = f_1$.

Let α = the value of ϕ at which $X = X_o$ then

$$X_o = \frac{E_g R_p}{E_1 \sin \alpha}$$

Since f_2 was originally greater than f_1, then X_o must be negative (that is, capacitive), in order to lower the frequency of oscillation. Hence $\sin \alpha$ must be negative. There are two values that satisfy this condition, one in the third quadrant and one in the fourth. Let the third quadrant value be called α_1 and the fourth quadrant value α_2.

If ϕ lies in the first or second quadrants, X is positive and the rate of change of ϕ (that is, $2\pi\ (f_2 - f_1)$) is large.) If ϕ lies in the third quadrant, but is less than α, $2\pi\ (f_2 - f_1)$ is small, but is still positive; that is, ϕ increases slowly. If ϕ should happen to have a value greater than α_1 but less than α_2, then X is greater than X_o, and $2\pi\ (f_2 - f_1)$, will be negative, that is, ϕ will decrease. If ϕ is greater than α_2, then $2\pi\ (f_2 - f_1)$ is again positive in value until $\phi = \alpha_1 + 2\pi$

From the above it is readily seen that, regardless of its initial value, ϕ will approach α_1 and remain at that value. It may be shown by means of a similar argument that if f_2 is originally less than f_1, the rate of change of ϕ is negative when E_g and E_1 are in phase. ϕ approaches a stable value α_1 in the second quadrant. The point of unstable equilibrium (where $\phi = \alpha_2$) is in the first quadrant.

If the frequency to which the circuit itself tunes is increased above f_1, the angle α_1 approaches $\frac{3\pi}{2}$ in value. If the circuit is tuned to a still higher frequency, f_2 will always be greater than f_1 and the rate of change of ϕ is always positive. In this case a beat note will be heard.

If the circuit is tuned below f_1, α_1 approaches $-\frac{3\pi}{2}$ in value and again, if the mistuning is carried farther, a beat will occur.

On the other hand, as the circuit is tuned toward f_1, the value of α_1 approaches π or 180° while α_2 approaches 0. If the circuit is tuned fairly accurately to f_1, the amplitude of the signal E_1 may be diminished to a very small value and E_g will still remain

in approximately the correct phase relation, due to the stabilizing effect of X_o, which comes into play when the phase tends to shift. This is the well-known phenomena of zero beat reception. Here, as in heterodyne reception, an optimum initial value of E_g will be found by trial.

REGENERATION AND SHARPNESS OF TUNING

It is a well-known fact that regeneration sharpens the tuning of a circuit by amplifying resonant frequencies to a greater extent than non-resonant frequencies. An analysis of the amplification at non-resonant frequencies will enable us to construct the resonance curve for any given tuned regenerated circuit.

First, consider a few facts concerning a non-regenerated circuit as shown in Figure 3. For convenience we assume the frequency f_1 of the applied voltage E_1 to be varied and the resonant frequency f_2 to be constant. If f_1 is higher than f_2, the tuned circuit offers a capacity reactance and the current thru R_p is leading E_1. At this time E_g is leading E_1 by some angle ϕ. As f_1 approaches f_2, E_g approaches 180° out of phase with E_1. When f_1 is less than f_2, ϕ has increased to more than 180°, or E_g is lagging with respect to E_1.

As shown under the discussion on "Locking-In and Zero Beat Reception," the current which is 90° out of phase with E_g is

$$I_x = \frac{E_1 \sin \phi}{R_p}$$

The equivalent reactance shunted across the tuned circuit thru which this current is flowing is $X = \dfrac{E_g R_p}{E_1 \sin \phi}$. It was also shown that ϕ is of such value that X tunes the circuit to resonance with the applied signal. For simplicity, we assume in this discussion that the degree of regeneration is not changed by retuning with purely reactive shunts, and also that the tube used has straight line characteristics.

From the equation giving the value of X, the value of E_g may be obtained. $E_g = \dfrac{E_1 X \sin \phi}{R_p}$. Now let us regenerate the circuit by adding a tickler, giving the circuit of Figure 2, and adjust the tickler to obtain critical regeneration. At any instant $E_1 \cos \phi$ is the effective signal applied to the circuit. Since the circuit is now regenerated, E_g increases. However, as E_g increases, the angle ϕ approaches 90°, which increases $E_1 \sin \phi$. It is evident that the limit is the point where ϕ reaches 90° in value, for at that point $\sin \phi$ is maximum. Notice also that the

in-phase component of E_1 is zero. Substituting the maximum value of $sin\ \phi$ in the above equation, we get $E_g = \frac{E_1 X}{R_p}$, which is the value to which E_g will build. As f_1 approaches f_2, the value of X increases and E_g increases. The resonance curve obtained, plotted between E_g and the applied frequency f_1 is, therefore,

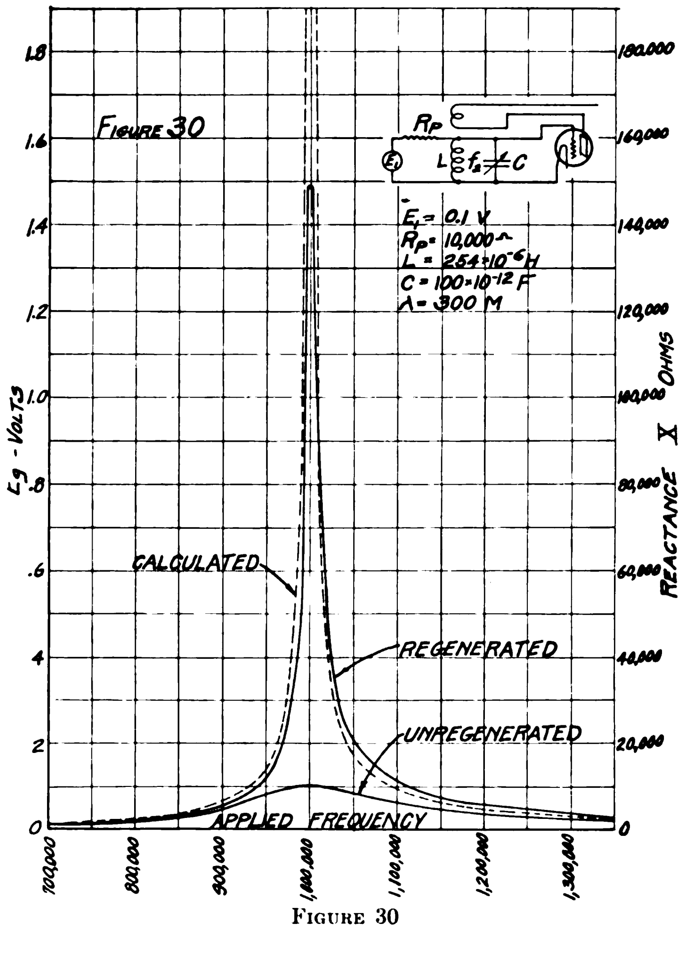


FIGURE 30

proportional to the curve plotted between f_1 and the parallel reactance necessary to tune the circuit to resonance at the frequency f_1. Such a resonance curve is shown by the dashed line in Figure 30.

The resonance curve obtained by the above method is modified by two factors. First, the limiting effect of the functional

terms of equation (4) prevent E_g from reaching an infinite value, the actual maximum value being that given by equation 4.

Second, critical regeneration is not maintained when the circuit of Figure 2 is retuned by reactive shunts. When an inductive reactance is shunted across a tuned circuit, the loss in that circuit *decreases.* (E_g constant). When a capacity reactance is shunted across such a tuned circuit, the loss *increases.* If adjusted for critical regeneration at the frequency f_2, and if f_1 applied is higher than f_2, the circuit is capable of oscillating at a frequency f_2. This is due to the decreased loss when the resulting inductive reactance is shunted across the circuit.

This tends to raise the amplification above that given by the above equation. Conversely, if f_1 is lower than f_2, the loss is increased and the amplification is less than that indicated by the equation. The effect of these variables on the curve derived from the equation is shown by the regenerated resonance curve in Figure 30.

Application to an Antenna

As will be observed, all of the above discussion has applied only to a regenerative stage of radio frequency amplification. When the same type of discussion is attempted concerning a regenerated antenna circuit, it is somewhat more difficult to draw an equivalent circuit. This is because it is difficult to show, in a simple way, the manner in which the signal voltage is applied.[2]

Whether the signal voltage is introduced into an antenna in series or in any other conceivable way the following remarks must hold true for it or for any type of regenerated circuit.

The equation expressing critical regeneration must take the form

$$\frac{E_g^2}{Z} = \frac{b\,E_g^2}{R_t}$$

where Z is the impedance of the tuned circuit, and E_g is an infinitesimal voltage.

When an amount of power "W" is introduced into the circuit or if "W" watts are prevented from being lost, there is a surplus of power which tends to charge the tuning condenser to a higher

[2]The authors believe that, mathematically, the signal voltage E_1 may be considered as being applied in series with a hypothetical resistance shunted between antenna and ground binding posts of the regenerated receiver; the resistance being of such a value as to replace the actual radiation resistance in damping effect. When this assumption is made, the regenerated antenna circuit reduces to the same equivalent circuit as the regenerated transformer, and the same formulas apply thruout.

value. As E_g increases, the tickler input can no longer be represented by $\frac{b\,E_g^2}{R_t}$ but is less by an amount $f_{bt}\,(E_g)$.

A condition of balance[3] will be obtained such that

$$\frac{E_g^2}{Z} = W + \frac{b\,E_g^2}{R_t} - f_{bt}\,(E_g).$$

If the circuit resistance is increased so that Z decreases, the tickler setting must be increased to maintain critical regeneration. If the tickler coupling is greater, the term f_{bt} (E_g) is larger for a given value of E_g. Hence the circuit comes to a balance with a smaller value of E_g than previously.

Since the term f_{bt} (E_g) is larger when a grid leak and condenser is used, the point of balance must always be lower with a grid leak and condenser.

In a similar manner, the principal points discussed with regard to a regenerated transformer could be applied to a regenerated antenna, and the conclusions resulting must be the same.

Conclusion

The most important point in a discussion of this kind is the character and point of application of the input voltage. It may be considered as a series voltage in the secondary circuit if it is recognized that the value of this voltage will vary with any change in the current in the secondary circuit. Thus the theory of voltage summation as cited in the introduction is unsound, even if the variations of tube characteristics are considered, the reason being that E_1 (at the point considered) will vary from cycle to cycle in any practical circuit. The only way any voltage may be considered as unaffected in value by load conditions is when the impedance of the generator is considered in the discussion.

In this paper the validity of the existing theories of regenerative amplification are questioned and an explanation offered based on a power balance. The circuit conditions affecting this power balance are discussed. The resulting amplification under various circuit conditions are given, including grid leak and condenser. Some of the more commonly observed phenomena are discussed in relation to the theory advanced in this paper. The time relations involved are briefly discussed and their effect considered. The entire theory may then be transferred to an antenna.

[3]For simplicity the terms f (E_g) and $f_g\,(E_g)$ are omitted in this general equation. Neither of these terms are very important compared to $f_{bt}\,(E_g)$.

In conclusion we may state that altho this paper seems rather long, many facts regarding regenerative amplification as affected by possible circuit changes are omitted. This was not due to lack of application of the present theory, but due to time limitations. Only such experimental proof as was necessary in direct confirmation of the theory was included.

It is hoped that in some measure this paper may serve to make this most useful phenomena of regeneration a little clearer and even more useful.

East Pittsburgh, Pennsylvania,
January 1, 1925.

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The authors are indebted to the following engineers of the Westinghouse Electric and Manufacturing Company for constructive criticism: Mr. L. W. Chubb, Mr. M. C. Batsel, Mr. C. R. Hanna, and Mr. Frank Falknor.

SUMMARY: This paper shows some of the defects of present theories regarding regeneration and presents a new method of analysis based on the idea of a power balance.

It is shown that a signal voltage does not supply power to a regenerated circuit but merely prevents certain losses from occurring. This upsets the balance between power input from the tickler and power lost in the circuit, so that oscillation occurs. In other words, regeneration consists of self-oscillation started and controlled by the signal voltage.

The amplification obtainable in this way has a definite limit, the limit

being caused by variations in the plate and grid impedances of the vacuum tube, as the amplitude of the grid voltage increases. The rate of variation of these impedances as the grid voltages increases, depends on the tube and on the direct voltage used.

The use of a grid leak and condenser decreases the voltage amplification, by increasing the rate of change of the plate filament impedance. In general, however, increased detecting efficiency more than makes up for the difference when audio frequency output is considered. The effect of resistance in the grid circuit is to decrease the amplification by increasing the effect of the impedance variations.

It is found that the best turn ratio to use in a regenerated transformer is the same ratio that should be used in a non-regenerated transformer. The amplification obtainable increases rapidly as the strength of an applied signal is decreased. Altho the inductance-capacity ratio does not affect the amplification obtained on an alternating current wave train, this ratio does affect the amplitude of the audio output when a modulated signal is being amplified. If a low L/C ratio is used, high notes will be lost when a weak signal is being received with full regeneration.

Regenerative amplification also occurs when a tube is in a condition of self-oscillation providing the strength of the local oscillation is weak. Under these conditions an incoming signal will be amplified whether it is beating with the local oscillation or of exactly the same frequency, as in zero beat reception. The dead spot noticed in the center of a beat note is caused by a retuning of the local circuit by the signal voltage. This voltage being out of phase with the grid voltage causes the phase of the current to shift, thus acting like a reactance.

A regenerated circuit amplifies non-resonant frequencies to a certain extent, the amount depending on the value of the reactance that would be needed to tune the circuit to the non-resonant frequency.

Altho the whole development is concerned with a regenerated transformer, the entire paper applies equally well to a regenerated antenna circuit.

DESIGNS AND EFFICIENCIES OF LARGE AIR CORE INDUCTANCES*

By

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(Radio Engineering Department, General Electric Company, Schenectady, New York)

In the operation of a radio transmitting station, two factors are of prime importance: reliability of operation and efficiency of operation. The antenna loading inductors, or tuning coils, may limit either of these two. In the early development of antenna tuning coils for use with 200-kilowatt Alexanderson alternators, reliability of operation was of first importance. With the problem of reliability solved, the question of efficiency became paramount. This paper describes, briefly, early designs which have proven to be reliable, and, in greater detail, recent designs which are reliable and more efficient.

The requirements for large air core inductances may be divided in two general classes: indoor coils and outdoor coils. In the majority of cases, outdoor coils are entirely satisfactory. In special cases in which the outdoor coils would be subjected to salt spray from the ocean, it is advantageous to protect the coils from the salt deposit. This requirement has lead to the development of an indoor coil. The designs of so-called indoor and outdoor types are essentially the same—the outdoor coil having longer creepage distances between turns and greater strength to withstand mechanical loads due to wind, and so on.

During the early development of tuning coils, wood which was used for the framework, even when protected from the weather, caused many fires and a necessary reduction in maximum antenna voltage and power. The next step forward was the substitution of porcelain for the wood, resulting in a design (Figure 1), which was superior to the wood in reliability, but which had metal supporting rings, thereby introducing additional eddy current losses. With the arrangement of the conductor on this coil, which formed a two-layer winding, the inductance and kilovolt-amperage capacity were relatively small for the physical

*Received by the Editor, April 29, 1925.

dimensions of the coil. To obtain greater inductance in the same length of coil, the multi-layer block (Figure 2) was developed. The coils using this type of block (Figures 3 and 4) not only have more turns per unit length of coil, but the spacing between turns is greater and the creepage distance many times greater.

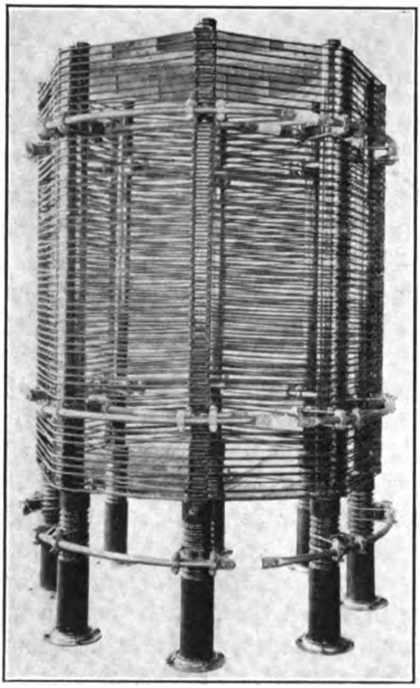

FIGURE 1—Antenna Tuning Coil

On a frame of given dimensions, it is possible to obtain with the same factor of safety approximately three times the kilovolt-amperage in the design shown in Figures 3 and 4 as in the design shown in Figure 1. By factor of safety is meant the ratio of the kilovolt-amperage at which creepage begins, to the operating kilovolt-amperage. Coils using this type of block may be wound in a variety of ways. The conductor may be wound in every groove, or as many grooves may be left idle between turns as is required for the voltage between turns. The same coil form may be wound as an inductor for low currents and high voltage or for low voltage and high currents. Small variations in inductance may be made by changing the spacing between turns. Two windings may be alternately interplaced in the grooves of the

blocks to form a transformer. The great variety of possible arrangements of conductors makes it possible to build a transformer to meet practically any requirement.

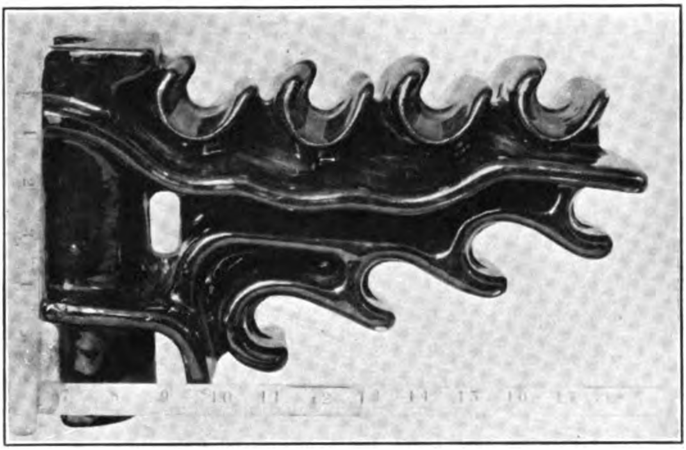

FIGURE 2—Porcelain Spacing Block for the Conductor of Antenna Tuning Coil

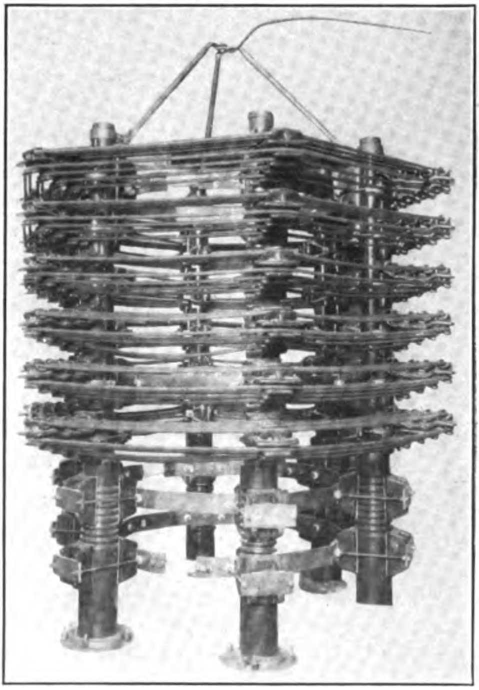

FIGURE 3—Tuning Coil for United Fruit Company, Radio Equipment

The construction of an outdoor type of coil, which has resulted from this development, is a porcelain braced coil, as shown in Figures 5, 6, and 7. This design eliminates almost all of the metal in the coil structure, and increases the rigidity of this beyond all former coils. The coil is made almost entirely of por-

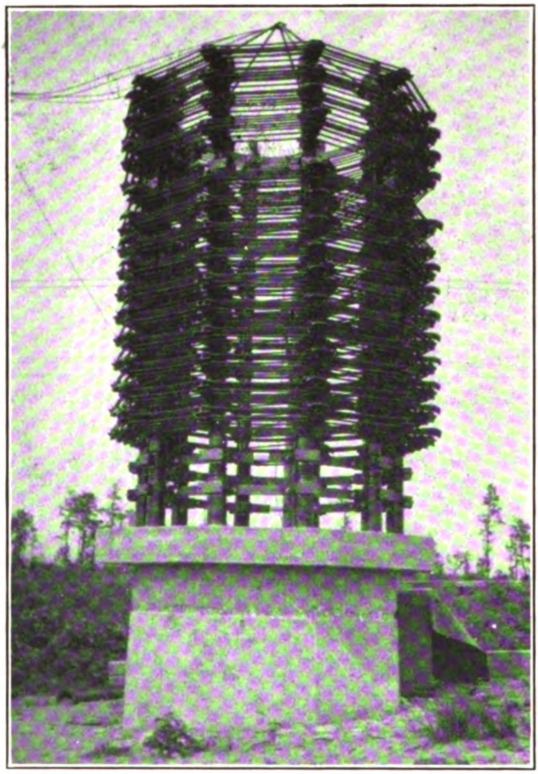

FIGURE 4

celain with the internal braces arranged to take advantage of the stiffness of a triangular form. During the development of this type of coil, a sample test coil was tested at the Tuckerton Station of the Radio Corporation of America. The umbrella antenna was being operated with one antenna coil in series with the alternator and in shunt with this circuit were three coils in multiple. Tests were made to determine the comparative efficiencies of the original coils, as shown in Figure 8, and the porcelain braced coil. In order to obtain data on the test coil under winter weather conditions, the conductor was wound in every other groove of the spacing blocks and the inductance

adjusted to the equivalent of the three shunt coils. The antenna resistance with this arrangement was lowered approximately sixteen percent from the value with the three-shunt coils. This coil was left in service three months, carrying approximately 18,000 kilovolt-amperes at 300 amperes during sleet, snow, rain, and fair weather conditions. As a result of these tests, the porcelain braced coil was shown to have higher efficiency at approximately four times the maximum kilovolt-amperage at which the original coils could be operated, and to be mechanically stronger than any previous types. In one case, the porcelain-braced coil was connected to replace one of the three multiple coils of the original type shown in Figure 1, and the reduction in antenna resistance was approximately 10 percent. As an extreme condition of loading, the test coil with conductor in every groove on

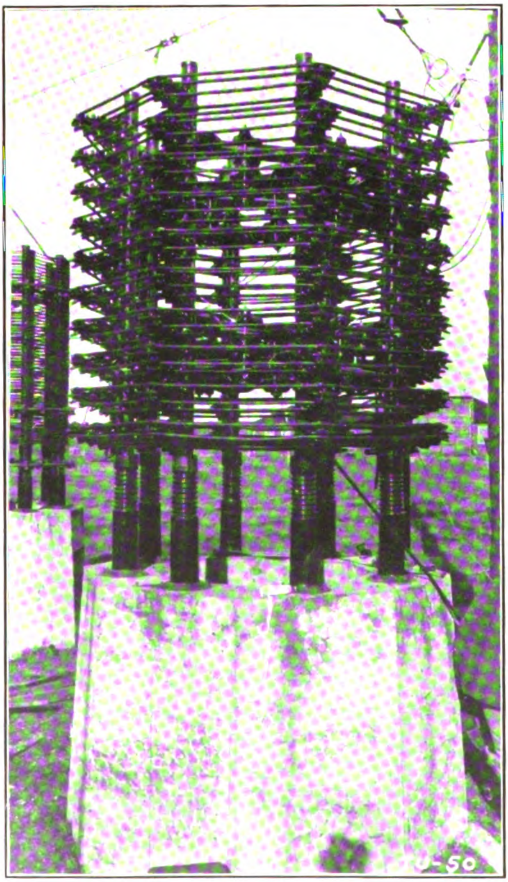

FIGURE 5

only four blocks, was loaded to more than 30,000 kilovolt-amperes at 400 amperes. This corresponded to an antenna input of approximately 200 kilowatts.

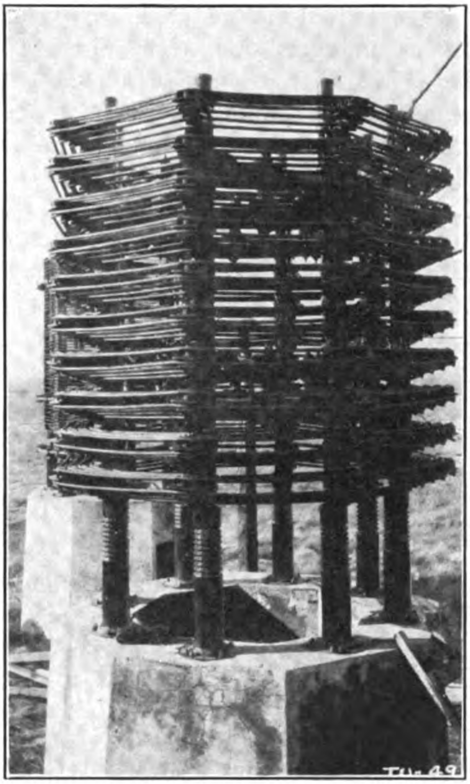

FIGURE 6

It is desirable to design a coil for indoor service more compactly than an outdoor coil for the following reasons:

1. Economy of space.
2. A concentrated field means less loss in surrounding objects.

The fact that shorter creepage distance for a given voltage provides the same factor of safety indoors as is obtained with longer creepage distances outdoors, makes it possible to utilize the compact design. With the smaller physical dimensions, the field, due to the coil, is concentrated, and the losses in any metal within the field would be greater than in a coil with a weaker field. For this reason, it is essential to eliminate all metal from the field except the conductor. Not only should all metal be removed, but the frame must be of good quality dielectric and be

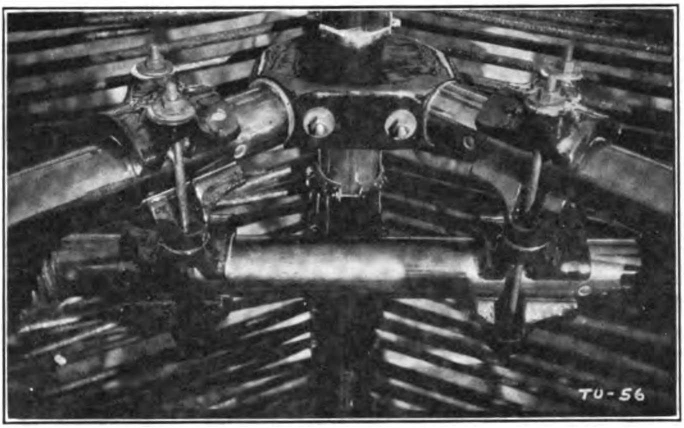

FIGURE 7

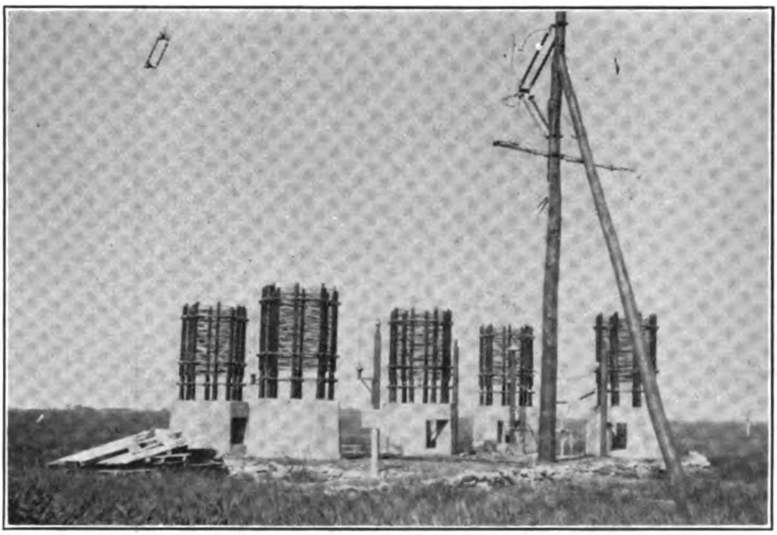

FIGURE 8

non-inflammable. The porcelain framework for such a coil is shown in Figure 9. In building up a coil of this type, three or four of the sections are stacked up, using mica sectors as cushions between the sections. This portion of the coil serves as a base and is formed of one to six sections, depending on the voltage, nearness to large metal objects, and the like. The active sections (Figure 10), may be wound before erection or may be set up as a continuation of the base and the spacing blocks and conductor applied afterward. The blocks (Figure 11) used on

this coil are similar to those used on the outdoor coils, but have been modified to provide a greater number of grooves in the same height. The spacing blocks are separated from the porcelain framework by mica cushions and are held in place by the conductor. This construction is very rugged: A man may stand on the end of a block without causing the conductor to sag or the block to slip. These sections are stacked in a vertical column until the required value of inductance is obtained. Each section is nine inches high, 33 inches inside diameter, and 75 inches outside diameter and has seventeen turns if all groves in the spacing blocks are filled. This coil meets all of the indoor requirements and, in addition, has an extremely flexible design. Only two porcelain parts are required.

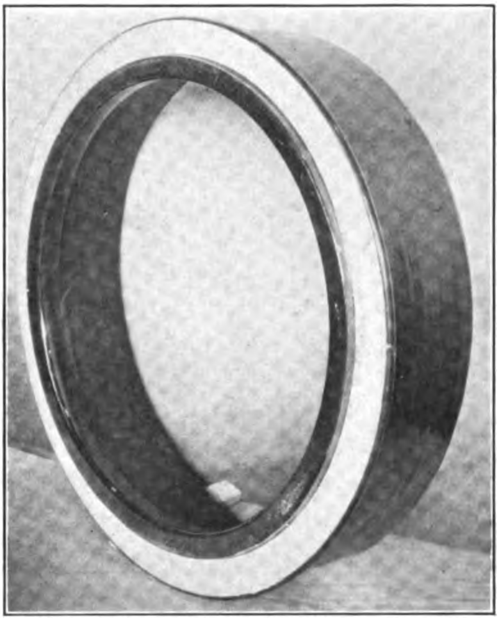

FIGURE 9—Porcelain Cylinder for Unit Type Tuning Coil 36 inches in Diameter, Bottom View

The efficiency of an antenna tuning coil is usually stated by giving the "power factor" of the coil. The power factor of an efficient coil should be below 0.002. Many of the coils in service are not within the range; however, there are some of the later designs which have power factors considerably below 0.001. Chief among the losses which go to make up the power factor of a coil are: losses in the coil frame, conductor eddy current loss,

and conductor ohmic loss. In the case of the indoor coil just described, the losses in the frame have been made extremely small so that the losses in the conductor are of the greater importance. The power factor due to the ohmic drop in the conductor is easily calculated from the direct current resistance and the reactive drop; that is

$$\text{Conductor Ohmic Power Factor} = \frac{R}{X}.$$

Mr. E. F. W. Alexanderson has developed a formula for the eddy current power factor. This formula was developed with the

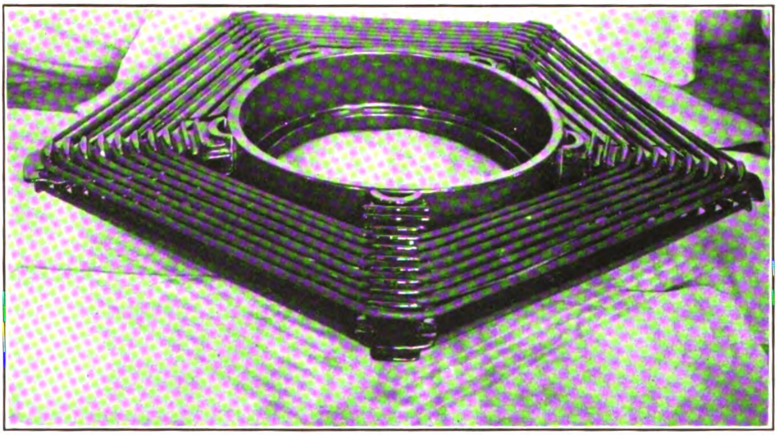

FIGURE 10—One Section of Unit Type High Power Antenna Tuning Coil

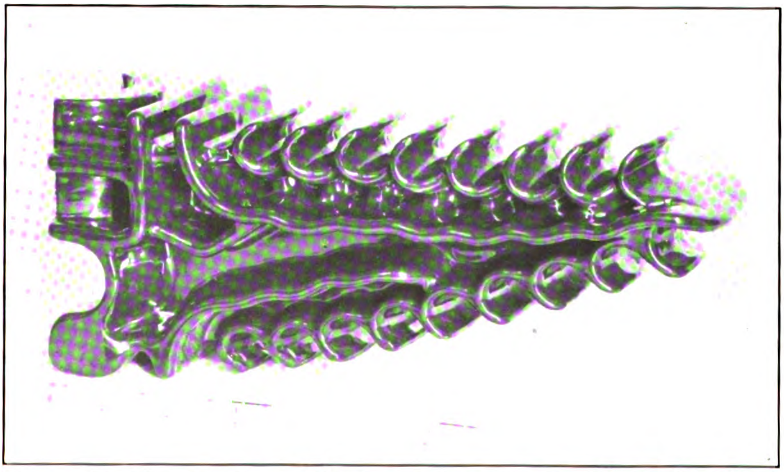

FIGURE 11—Porcelain Spacing Block with Seventeen Grooves, for Unit Type High Power Antenna Tuning Coil. Side View

assumption that the conductor was in a uniform magnetic field. This field is not absolutely obtained but is very closely approximated in the larger coils. This formula

$$\text{Conductor Eddy Current Power Factor} = \frac{X}{R}\left(\frac{d}{D}\right)^2$$

does not take into account the form of the conductor; that is, whether there is a non-conducting core or the size of such core. In the preceding formulas:

X = Reactance
R = Direct current resistance
d = Diameter of individual strands
D = Coil diameter

Tests which have been made indicate that this formula is correct to within a few percent for all conductors commonly used in high power coils. The calculated ohmic and eddy current power factors of the unit type coil with four different conductors is shown in Figure 12. The eddy current power factor increases directly with the frequency, while the ohmic decreases inversely with the frequency. The small circles at the intersection of the ohmic and eddy current power factor for each conductor indicate the optimum frequency for the coil with that particular conductor. Figure 13 shows the sum of the ohmic and eddy current power factors on the conductor number 2 with various diameters

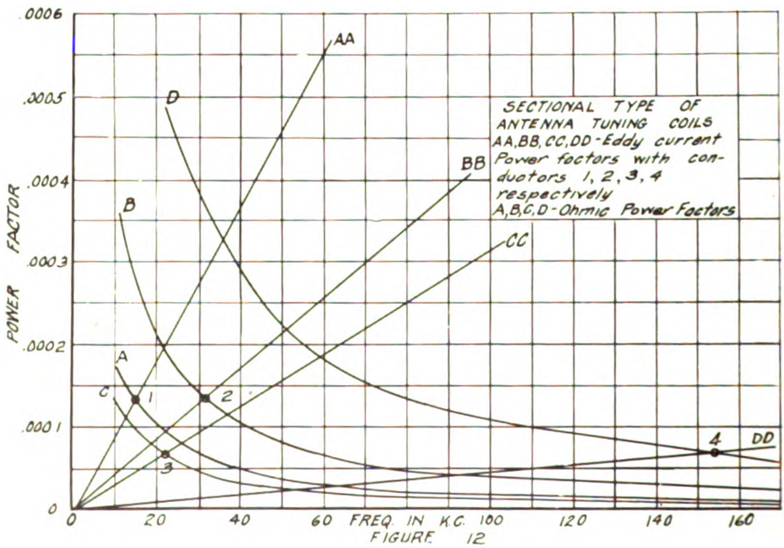

FIGURE 12

of coil. The increased efficiency of the 75-inch diameter over the 60-inch is very marked, but the increase from 75 inches to 120 inches is not so great. For this reason, 75 inches seems to be a good diameter for an indoor coil, using this conductor.

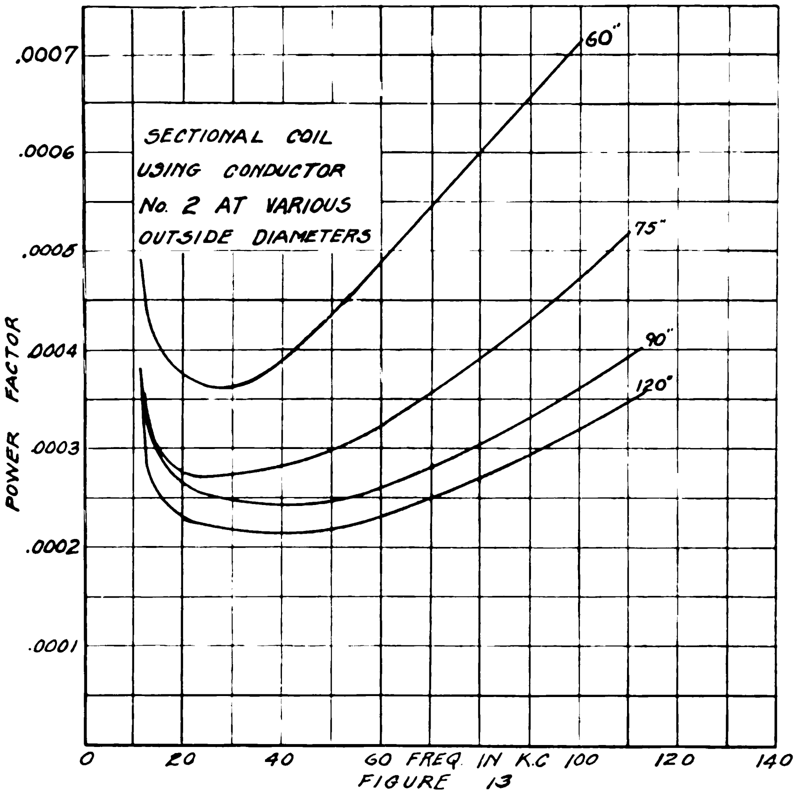

FIGURE 13

For the large air core inductors which have been described, the conductors are all of the Litzendraht or finely stranded copper conductor, each strand enameled. A large number of tests have been made on the conductors for this use. These tests show that the smaller the diameter of the strands and the larger the non-conducting core, the lower the losses are in the conductor.

March 31, 1925,
Schenectady Works, New York.

SUMMARY: Representative designs of large air core antenna tuning inductances are described, in types suitable for outdoor and indoor service. The latest improved designs are described in greater detail and compared with earlier designs on a basis of efficiency and kilovolt-ampere capacity. Formulas

are given for calculating the ohmic and eddy current conductor power factor of coils using finely stranded, separately-insulated strands.

In graphical form are shown the variations of ohmic and eddy current power factor with frequency, with four different conductors wound in a given arrangement to given dimensions; also the variation of the sum of ohmic and eddy current power factors with frequencies for a representative conductor on various diameters. These values were calculated by the formulas given, and indicate very high efficiencies for the latest types of coils.

AN EFFICIENT TUNED RADIO-FREQUENCY TRANSFORMER*

By

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At a time when radio engineers are generally concerned with the problem of developing more sensitive methods of reception, it seems particularly desirable that exact data on the design of an efficient tuned radio-frequency transformer be made generally available. While the present paper deals specifically with the problem of radio-frequency amplification over the broadcast range, some general formulas are furnished for the design of a tuned radio transformer for any wave length band.

The figure of merit of a radio-frequency amplifier is usually considered to be its voltage amplification factor, but the ability of the amplifier to discriminate between the desired signal and signals at other frequencies is a factor of considerable importance in determining its utility. In preparing this paper, it has been our aim, first to design a transformer so that the theoretical amplification could be obtained, then to consider its selectivity, and the tendency of the circuits to break into oscillations. Values of voltage amplification predicted by theory have been checked by experiment, and from this agreement, as well as from audibility tests, the realization of the theoretical resonance curve has been concluded.

The radio-frequency transformer is represented in Figure 1 by coils L_1 and L_2, having a mutual inductance M. The primary of the transformer is untuned; the secondary is tuned by a condenser C_2. Since the primary coil L_1 is in series with the plate resistance of a vacuum tube, its resistance may be neglected in computing the current in the secondary. This fact, as will subsequently be shown, is of importance in designing the primary so that capacity coupling with the secondary will be negligible.

In this paper the over-all voltage amplification of tube and

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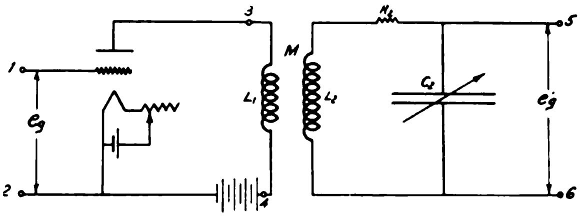

FIGURE 1

transformer is called K, and is equal to e'_g/e_g. (See Figure 1.) In comparing the values given for K with the voltage amplification obtained from a one-stage audio-amplifier, it should be remembered that the response of a vacuum tube detector is nearly proportional, for small values of e_g', to the square of the applied variation voltage so that a value $K=10$ is equivalent to an audio-amplification of 100. (A good one-stage audio-amplifier will give a voltage amplification of 30.)

It may be shown that for alternating current, Figure 1 is electrically equivalent to Figure 2 where the amplifier tube is replaced

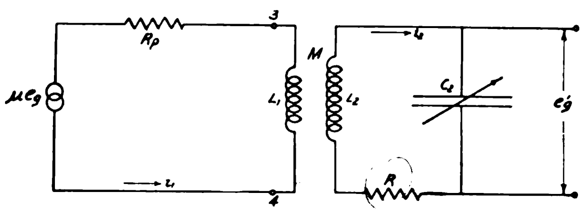

FIGURE 2

by an emf. of $\mu\, e_g$ in series with the plate resistance R_p, and the problem of determining the value of K is reduced to the problem of solving Figure 2 for i_2, since to an extremely close approximation

$$e'_g = L_2\,\omega\, i_2 \tag{1}$$

The general coupled circuit expression for secondary current is[1]

$$i_2 = \frac{M\,\omega\,\mu\, e_g}{Z_2\, Z_1'} \tag{2}$$

where
$$Z_2^2 = R_2^2 + X_2^2$$

$$(Z_1')^2 = \left(R_p + \frac{M^2\,\omega^2}{Z_2^2} R_2\right)^2 + \left(X_1 - \frac{M^2\,\omega^2}{Z_2^2} X_2\right)^2$$

The secondary current may be expressed in a form better

[1] Pierce, "Electric Oscillations and Electric Waves," Chapter XI.

adapted to computation and design by employing certain standard abbreviarions, as follows:

$$\tau = \frac{M}{\sqrt{L_1 L_2}}; \quad \gamma_1 = \frac{R_p}{L_1 \omega}; \quad \gamma_2 = \frac{R_2}{L_2 \omega}; \quad \frac{\omega}{2\pi} = f\begin{pmatrix}\text{impressed}\\ \text{frequency}\end{pmatrix}$$

$$X_1 = L_1 \omega; \quad X_2 = L_2\omega - \frac{1}{C_2 \omega}; \quad \omega_o = \frac{1}{\sqrt{L_2 C_2}}; \quad \frac{\omega_o}{\omega} = \theta$$

μ = amplification factor of the tube.

With this notation and equations (1) and (2), the over-all voltage amplification factor becomes

$$K = \frac{e_g'}{e_g} = \frac{\mu \tau \sqrt{\dfrac{L_2}{L_1}}}{\sqrt{\gamma_2^2 + (1-\theta^2)^2}\sqrt{\left\{\gamma_1 + \dfrac{\tau^2 \gamma_2}{\gamma_2^2 + (1-\theta^2)^2}\right\}^2 + \left\{1 - \dfrac{\tau^2(1-\theta^2)}{\gamma_2^2 + (1-\theta^2)^2}\right\}^2}} \tag{3}$$

In equation (3), τ is the coefficient of coupling between coils L_1 and L_2 and γ_1 and γ_2 are the power ratios of primary and secondary circuits. It should be noted that γ_2 is the usual sort of power ratio for a coil and will be nearly constant over a considerable range away from the natural period of the coil, while γ_1 contains a resistance independent of the frequency and cannot be considered constant over any range.

In tuning the transformer for maximum signal strength, condenser C_2 is varied until i_2 is a maximum, assuming, of course, that L_1 and L_2 are fixed. This partial secondary resonance relation is expressed mathematically by the relation $\dfrac{X_2}{X_1} = \dfrac{M^2 \omega^2}{Z_1^2}$, which,[2] using the notation above and equation (2) gives

$$K_{max} = \frac{\mu \tau \sqrt{\gamma_1^2 + 1}\sqrt{\dfrac{L_2}{L_1}}}{\gamma_2(\gamma_1^2 + 1) + \tau^2 \gamma_1} = (\text{approx.}) \frac{\mu \tau \sqrt{\dfrac{L_2}{L_1}}}{\tau^2 + \gamma_1 \gamma_2} \text{ if } 1 \ll \gamma_1^2 \tag{4}$$

So far no special values of the transformer constants have been assumed except that L_1 and L_2 are fixed inductances. From equation (4) it is evident that since γ_2 is approximately constant, there will exist a relation between τ and γ_1 which will make K a maximum for a particular frequency. This relation may be shown to be

$$\tau^2 = \frac{\gamma_2}{\gamma_1}(\gamma_1^2 + 1) = \text{approx. } \gamma_1 \gamma_2 \text{ as } 1 \ll \gamma_1^2 \tag{5}$$

[2] Previous citation.

Equation (5) can only be satisfied for one frequency. For this one frequency the optimum K will be,

$$K_{opt}=\frac{\mu\sqrt{L_2/L_1}}{2\sqrt{\gamma_1\gamma_2}}=\frac{\mu L_2\omega}{2\sqrt{R_p R_2}} \tag{6}$$

In designing the transformer, it is, in the light of the (6), clearly desirable to make L_2 as large as possible consistent with ability to tune to the minimum wave length desired. The actual value of L_2 is thus a function of the distributed capacity of L_2, of the minimum capacity of C_2, and of the input capacity of the next tube. The value of frequency for which $\tau^2=\frac{\gamma_2}{\gamma_1}(\gamma_1^2+1)$ may conveniently be chosen in the middle of the wave length band for which the transformer is designed.

For plotting curves of the various equations above, the following constants have been chosen. These represent about the best design for a tuned radio-frequency transformer to include wave lengths between 300 and 600 meters with UV-199 tube. Other tubes having equally small input capacities will serve as well.

$L_1=0.13$ mh.	$R_p=20{,}000$	$\mu=6$
$L_2=0.400$ mh.	$\gamma_2=0.008$	$\tau=0.52$

With these constants $\tau^2=\frac{\gamma_2}{\gamma_1}(\gamma_1^2+1)$ at 400 meters.

Figure 3 shows a curve of K against τ for different values of wave length. These curves are plotted from equation (4) with n_2 assumed constant, and show that so long as τ is greater than 0.4, good amplification will be obtained over the broadcast range. The maxima of these curves are the points where $\tau^2=\frac{\gamma_2}{\gamma_1}(\gamma_2^2+1)$ and the dotted line is, therefore, the locus of that equation. Figure 4 is the value of γ_2 computed from laboratory measurements of R_2. This R_2 was measured with a lighted UV-200 tube, connected across the transformer secondary. L_2 consisted of 70 turns of number 20 double cotton-covered wire wound on a tube four inches in diameter, while the secondary condenser was one of special construction designed for low losses. The values of γ_2 given in Figure 4 are believed to be about the best that can be obtained with a reasonable size of L_2.

In order that the theoretical amplification could be obtained, it was early realized that capacity coupling between primary and secondary must be reduced to a value negligible compared with the inductive coupling. At the same

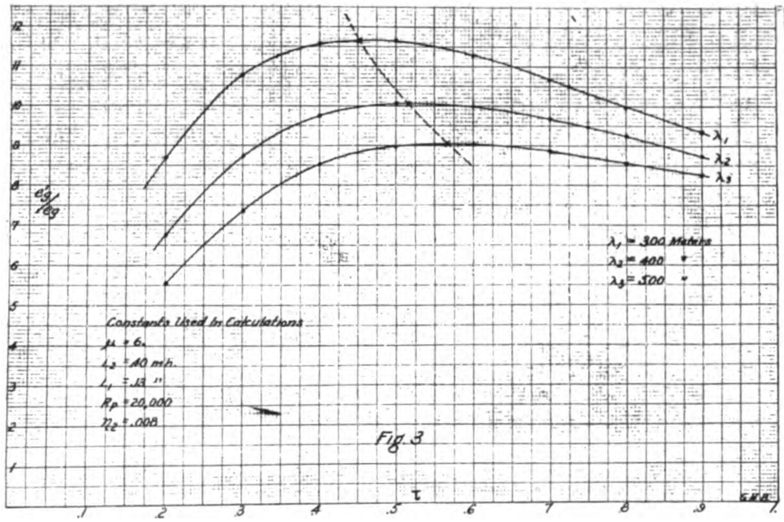

FIGURE 3—Amplification of a UV-199 Tube and Radio Frequency Transformer

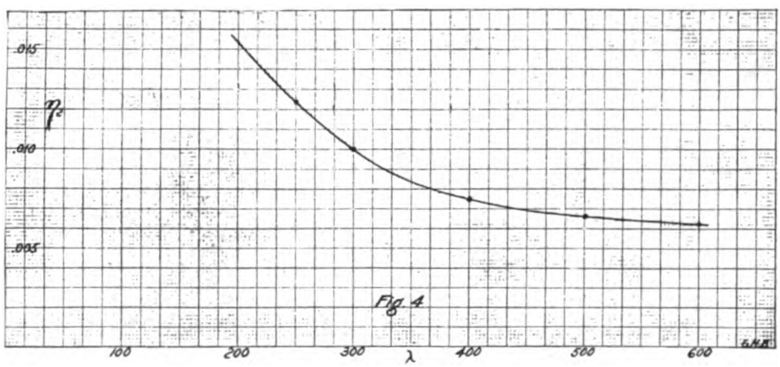

FIGURE 4—Curve of r_2 Against λ. Coil Coil Consisted of 70 Turns of Number 20 D.C.C. Wire on 4-Inch Tube. (Tube and Socket Included)

time, for reasons given later, as large a value of coefficient of coupling as possible is desirable. Capacity coupling may be assumed to be present when, in practice, a reversal of the connections to the primary of the transformer changes the value of secondary current. In general, capacity coupling will be very large if L_1 and L_2 are coaxial, single layer coils of slightly different radii. Figure 5 shows the design finally adopted for the transformer for use in the 300-600 meter range. The secondary is wound with number 20 double cotton-covered wire on a four-inch tube; the primary is wound in a small channel in a thin wooden cylinder tightly fitted inside the secondary

tube. Since the resistance of the primary is unimportant, the size of the wire for it should be chosen as small as can be readily handled and should be double cotton-covered to reduce distributed capacity in the primary itself. Experiments showed that such a transformer has very small capacity coupling, so that the primary may be placed in the center of the secondary where the inductive coupling coefficient is a maximum. Although the calculations and measurements were made with a 70-turn secondary, the value 65 turns given on Figure 5 is somewhat more suitable if signals from 240 to 300 meters are to be received efficiently. However, with a condenser C_2 of maximum capacity 0.0005 mf., and small minimum capacity, either 65 or 70 turns on the secondary will give sufficient inductance for reception on wave lengths from 240 to 600 meters.

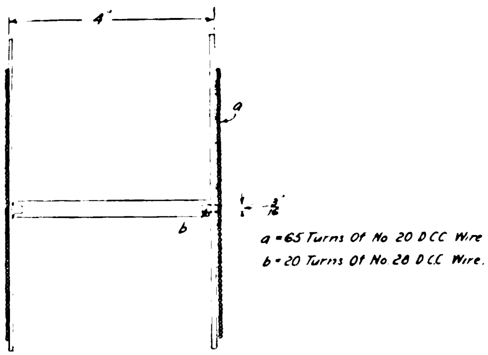

FIGURE 5

Having decided upon the transformer design, the next problem is to investigate its performance in an actual radio circuit. The possibility of creating sustained oscillations in a tuned circuit connected to the grid of a three-electrode tube by adding positive reactance in the plate circuit has been fully treated elsewhere.[3] A very rough idea of the amount of plate reactance necessary to produce such sustained oscillations may be formed from an examination of Figure 7, which was obtained by the method of Figure 8 for values of $L_o/C_o R_o$ thought to be average. The important fact about Figure 7 is the order of magnitude of the quantities involved.

Referring to Figure 1, it is readily seen that if an oscillatory circuit is connected between points 1 and 2 in suitable manner, oscillations may be produced due to the presence in the plate

[3] Stuart Ballantine, "Physical Review," Volume XV, May, 1920.

circuit, between points 3 and 4, of an "equivalent reactance," made up of the reactance of coil L_1 and the reactance produced by the secondary current thru M on L_1. The equivalent reactance in the plate circuit of the amplifier tube may be defined as the reactance, which, when connected between points 3 and 4 (Figure 1), would give the same value of alternating current between these points as the arrangement shown. The value of this equivalent reactance is given below equation (2), and by virtue of earlier notation is

$$X_1' = L_1 \omega \left\{ 1 - \frac{\tau^2 (1-\theta^2)}{\gamma_2^2 + (1-\theta^2)^2} \right\} \tag{7}$$

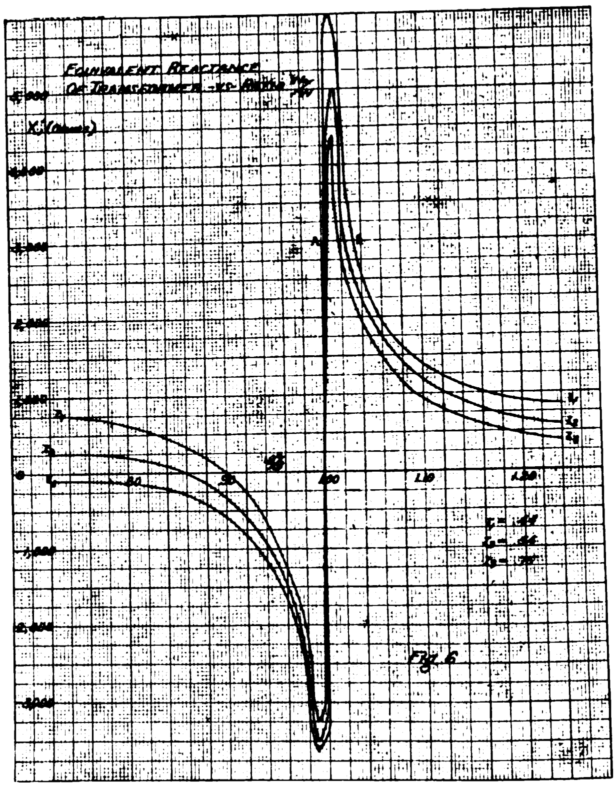

FIGURE 6

Figure 6 is plot of X_1' against θ for three values of the coefficient of coupling, for which $\tau^2 = \gamma_1 \gamma_2$. For large values of τ the value of L_1 is, therefore, small, and conversely. For the conditions of Figure 7 oscillations would only occur in a narrow region

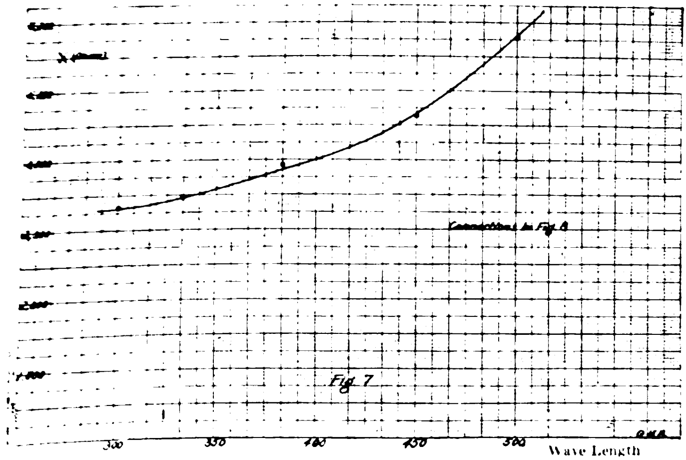

FIGURE 7—Reactance in Plate Circuit Necessary to Make a UV-199 Tube Oscillate

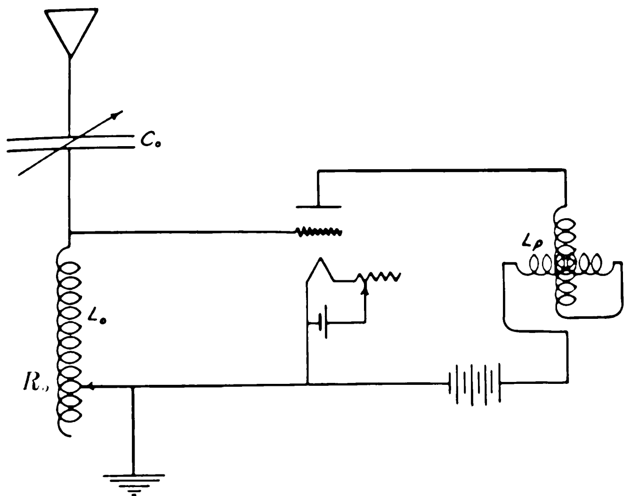

FIGURE 8

such as AB, and it is evident that AB will be narrower the greater the coefficient of coupling. The reason for making τ as large as possible, consistent with negligible capacity coupling, is now evident, for by doing so the region over which the amplifier will tend to oscillate is reduced to a very narrow band adjacent[4] to the setting for maximum voltage amplification, and by changing the value of L_o, C_o R_o (see Figure 10) by any of the conventional

[4] The exact value of θ for maximum amplification may be shown mathematically to be very slightly less than unity. At this point the reactance is positive but less than half its maximum positive value.

methods, the narrow band may be reduced to a point, thus providing a beautiful method of determining when the antenna is in tune with the radio amplifier. Once the operator has calibrated the amplifier condenser for wave length he has only to tune the antenna circuit until the narrow band is found over which the tube oscillates, and then to reduce this band to a point, after which the whole apparatus will be almost exactly in tune with the desired wave length. This process takes longer to describe than to perform.

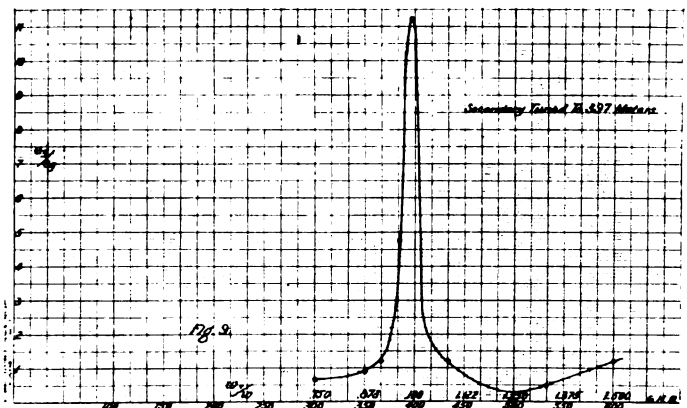

FIGURE 9—Resonance Curve of Transformer and UV-199 Tube. (Amplification Against Wave Length)

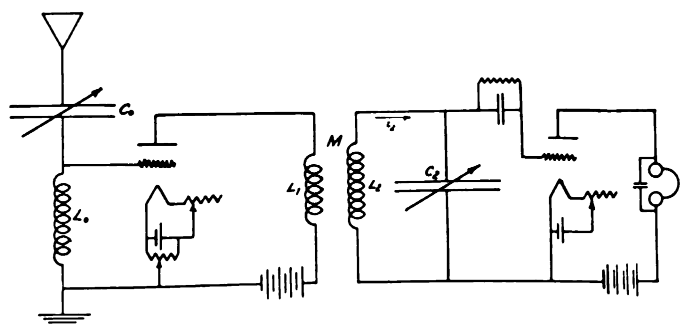

FIGURE 10

Figure 10 shows the transformer used in a single circuit one-stage radio-amplifier receiver with non-regenerative detector. A potentiometer serves to change the value of L_o/C_oR_o until the correct amount of regeneration is obtained. For the transformer above described used in conjunction with UV-199 tube, oscillations can be stopped before the grid becomes positive with respect

to the filament, so that no amplification will be lost by having the grid positive. This is possible because τ is chosen large and L_1 small, thereby satisfying the condition for maximum amplification and reducing to a minimum the tendency to oscillate.

The selectivity of a one-stage radio-amplifier using a UV-199 tube and the transformer here described is *relatively* very great. Equation (2) is the analytical expression for a resonance curve of K against wave length. Figure 9 is a plot of such a resonance curve for $\theta = 1$ at 400 meters, this point also being that at which $\tau^2 = Z_1 Z_2$. Resonance curves at other wave lengths will be of the same general character, but will have slightly lower maxima. The selectivity is said to be "relatively" great because when the transformer is tuned to the desired signal, other signals at nearby wave lengths are amplified to a much smaller degree, yet it is to be noted that there is practically a one-to-one voltage transfer at wave lengths thruout the broadcast band. The practical importance of this is that a strong local station will be heard when the transformer is tuned to another wave just as loudly as if the radio-amplifier tube and transformer were removed and the incoming voltage applied directly to the detector, but as soon as the weak station to which the transformer is tuned is picked up, its signals will experience an absolute amplification of about ten, and so a relative amplification of ten-to-one is obtained. If absolute selectivity is desired, a coupled circuit should be used before the radio-amplifier.

Experimental Verification of Amplification

The method used to measure the voltage amplification is believed to be somewhat novel, in that it depends essentially on extrapolation from curves for different conditions than those under which the transformer is used. Figure 11 shows the connections. The voltage applied to the grid-filament of the amplifier tube is known[5] from values of I and R. The current produced by this voltage in the secondary circuit is measured by a thermo-couple and galvanometer, T and M. Since the thermo-couple happened to have a resistance (1.5 ohms) nearly equal to that introduced in circuit $L_2 C_2$ by connecting the detector tubes to points 7 and 8, and since the total secondary resistance is about 20 ohms, no correction need be applied for

[5] Great precautions against stray fields must be taken. In these measurements twisted leads carried the radio-frequency current to the resistance from an oscillator located in another room.

the resistance of the couple, and the voltage amplification is given by

$$K = \frac{L_2 \omega i_2}{R I}$$

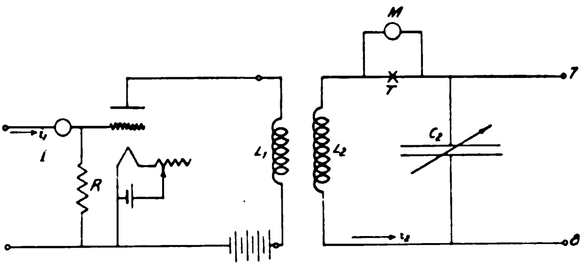

FIGURE 11

The most sensitive low resistance thermo-couple available would not give accurately measurable values of i_2 until the voltage applied to the grid of the first tube was about four volts, a value enormously greater than that produced by a strong nearby station in a tuned antenna and, therefore, available as the input voltage to the amplifier tube. Also, even with four volts on the grid, the value of i_2 could not be read with absolute accuracy, so for these reasons it was not thought desirable to rely wholly on a single value at such a large grid voltage, and the idea was conceived of obtaining several values of i_2 at grid voltages ranging from 4 to 15 volts and extrapolating to find the real i_2 for a value of grid voltage for which the excursion on the tube characteristic would not exceed the linear region. A series of such curves is shown in Figure 13. The value of plate voltage used was 50 volts. As long as μe_g is less than about 20, there is a linear relation between i_2 and e_g and a common value of four volts may conveniently be used in reading the corresponding values of i_2.

If the agreement of theoretical and experimental values for voltage amplification is taken as an indication of the validity of the experimental determination, the method outlined above is a valuable one, as Figure 12 shows. At the upper end of the broadcast band the agreement is almost perfect while divergence near the lower end is probably due to the increasing importance of small capacities in the tube and transformer at the higher frequencies and possibly due also to the fact that the value of R_p chosen to design the transformer was not that of the tube used

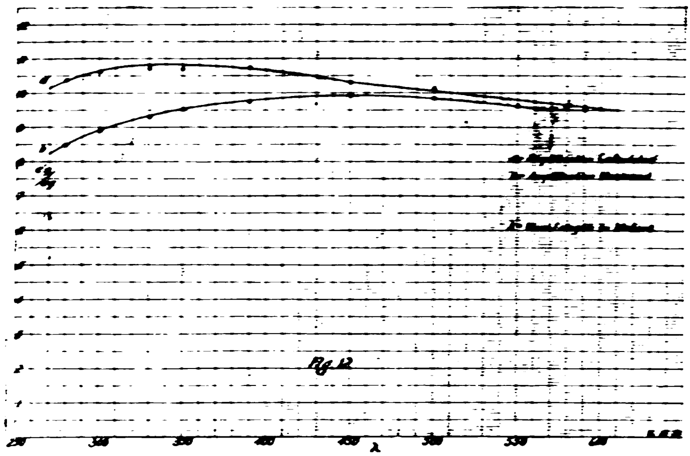

FIGURE 12—Amplification, Theoretical and Practical

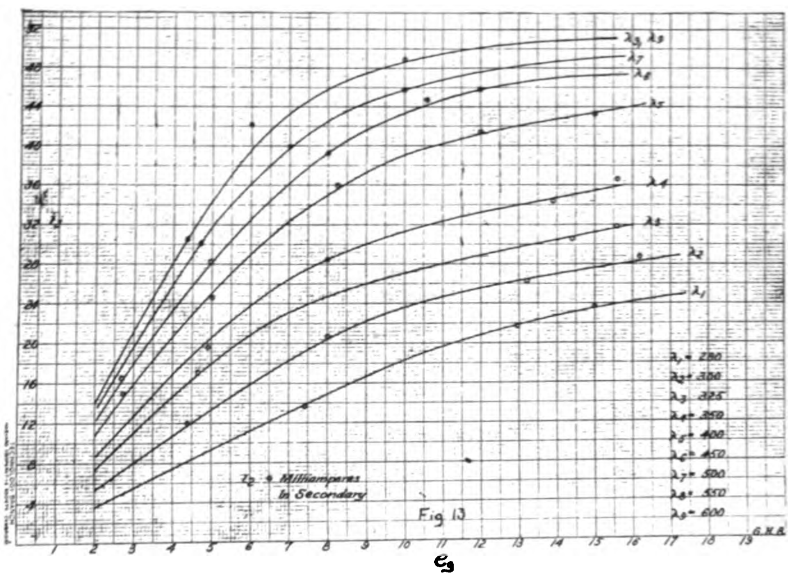

FIGURE 13—Current in the Secondary of the Radio Transformer Against Impressed Signal

in the measurements, so that the relation $\tau^2 = \gamma_1 \gamma_2$ was really satisfied by the transformer at a higher wave length than 400 meters. It is believed that those who have had experience with the difficulties of measurements of this sort will be lenient toward the slight discrepancies between the theoretical and experimental curves.

There are two good methods of obtaining radio amplification at short waves. One is that described in this paper and consists essentially in tuning the transformer to the signal. The other is that originated by Armstrong, the super-heterodyne, which tunes the signal to the transformer. The writers sincerely hope that the material here presented will prove of considerable value to those desirous of improving upon the sensitiveness and selectivity of the ordinary regenerative receiver by the former method. This paper should, therefore, appeal to the enthusiasts who are interested in radio-frequency amplification as well as to those who believe the utmost sensitivity is obtained by a combination of regeneration and tuned radio-frequency amplification.

SUMMARY OF DESIGN FORMULAS

1. To compute L_2, let λ_o be the minimum wave length in meters to which it is desired to tune. Then L_2 in millihenrys will be given by

$$L_2 = 5 \times 10^{-6} \lambda_o^2$$

2. To satisfy the relation $\tau^2 = \eta_1 \eta_2$, assume $\tau = 0.5$ (for the construction shown in Figure 5) and substitute the value of η_2. Then knowing the plate resistance of the amplifier tube and the mean frequency to which it is desired to tune, the value of L_1 can be computed.

3. For C_2 select a low loss condenser of proper capacity to cover the wave length band desired.

DISCUSSION*

ON

"A NEW PHENOMENON IN SUNSET RADIO DIRECTION VARIATIONS" BY L. W. AUSTIN

BY

R. L. SMITH-ROSE AND R. H. BARFIELD

(RADIO RESEARCH BOARD STATION, SLOUGH, BUCKS, ENGLAND)

The experimental results described by Dr. Austin in the August issue of the PROCEEDINGS are of great interest to us since we have been using tilting frame coil direction-finders for the past two years in an investigation of radio wave-fronts. As far as observations on long-wave stations at various distances are concerned, however, we have so far been unable to obtain any of the effects described by Dr. Austin, and we should like to ask one or two questions in order to obtain more definite information on this point. In a paper published earlier in the year (R. L. Smith-Rose and R. H. Barfield, "On the Determination of the Forces in Wireless Waves at the Earth's Surface"—Proc. Roy. Soc., 1925, Volume 107, pages 587-601), experiments were described in which the conductivity of the earth was measured at radio frequencies. In the same paper it was shown that this conductivity is so high in England that on long wave lengths, even were a radio wave arriving at the earth at an appreciable angle of incidence and polarized with its magnetic field in a vertical plane, the reflected wave set up at the earth's surface will be such as to tend to eliminate any vertical component of the magnetic field. The results of experiments were given supporting this theoretical prediction, and showed that the magnetic field of the arriving wave is always horizontal to within 1°, which was the limit of accuracy of the apparatus then employed. As a specific instance we may quote the observations on Leafield's transmission on a wave length of 12,400 meters at a distance of about 77 km. This station has been regularly watched now for over a year and altho all the usual effects of variations of bearings and signal intensity are observed, the resultant magnetic field has never departed from the horizontal by more than the limit above mentioned. Consequently, when the signal minimum with the coil vertical is blurred, indicating an elliptically-rotating magnetic field in the horizontal plane, no sharp signal minimum

* Received by the Editor, September 9, 1925

is found in any position except that in which the plane of the coil is swung thru the horizontal position, that is, with the vertical axis tilted at 90°.

In our own experiments somewhat elaborate precautions have been taken to eliminate any stray emf's which may be introduced into the receiving system by antenna or capacity effects or as a direct pick-up of signal on parts of the apparatus other than the frame coil. Furthermore, the site chosen for our experiments (at Slough) is situated well away from buildings, overhead or underground metalwork and so on, which might cause distortion of arriving waves. The site is known not to be ideal, but as it has been the subject of detailed investigation for the past four years, its essential defects are now fairly well understood. We should like to be assured that Dr. Austin's experiments have been carried out under similar conditions, for since the vertical component of magnetic field is at best only a residual effect, it might be possible to get comparatively large tilts by balancing the effect of this magnetic field against a stray emf. in some other part of the apparatus. On a site which is not free from metallic frame buildings or even hills and other geographical features, there is the further possibility that the arrival of downcoming waves at night will produce abnormal effects in the immediate neighborhood of such sources of disturbance. If, as we hope, the experiments are above such criticism, it is evident that the site at which they are made possesses a fairly low effective conductivity much lower than we have as yet been able to find in England. Could Dr. Austin give the results of any measurements of this conductivity at radio frequencies, either on the particular site in question or in any other parts of America? Having failed to find a place in this country (England) where the conductivity has a favorable low value, we have been compelled to resort to the alternative of using shorter waves in order to obtain measurable departures of the electric and magnetic fields at night from their normal day-time directions. Some measurements which have been carried out during the past three months on wave lengths between 300 and 500 meters have given very interesting results, and it is hoped to publish these shortly.

August 25, 1925.

L. W. Austin (by letter):* Of course I am very glad to find that Dr. Smith-Rose and Mr. Barfield have been interesting themselves in the phenomena described in my paper in the August

* Received by the Editor, October 29, 1925.

PROCEEDINGS, even though their observations do not agree with mine.

The conditions of experiment here were as follows: The double-axis radio compass coil and the well-shielded receiver were housed in a portable wooden garage about two hundred feet from the one-story radio building and a little nearer one of the 125-ft. steel towers which support the main laboratory antenna. The soil around the garage is a coarse gravel mixed with red clay and there is an average slope of the ground toward the southeast amounting to 8 or 10 degrees. Not far away there are a number of small trees. The radio-frequency resistance of the ground has never been measured with any accuracy, but from antenna resistance measurements it is believed to be high.

These conditions are not ideal, but special experiments show that the steel tower, which is by far the most dangerous source of error, plays no part in the phenomena. In these experiments the main antenna, supported by the tower in question, was tuned to WII, the station being observed for direction variation or blurred minimum, and alternately connected to the tuning inductances and left open. Within the accuracy of observation this produced no change in the deviation of the radio compass or in the vertical angle for which the minimum became sharp. The accuracy of setting was certainly better than 1 degree for the deviation and 5 degrees for the vertical angle.

Another experiment, in which the horizontal compass axis was pointed toward the transmitting station, showed a minimum of signal with the coil depressed about 10 degrees toward the southeast, i. e., bringing the coil roughly parallel with the sloping ground.

Still other experinents were made in which the antenna effect of the compass coil was varied by using a compensating capacity to earth and also by coupling a small untuned antenna to the coil. None of these, however, produced any certain changes in the phenomena.

Bureau of Standards, Washington, D. C.
October 26, 1925.

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DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

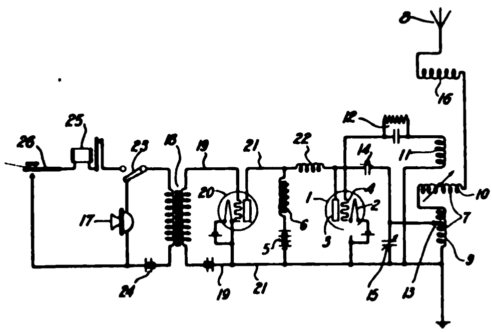
ISSUED SEPTEMBER 1, 1925–OCTOBER 27, 1925

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,551,624—J. C. Schelleng, filed June 30, 1924, issued September 1, 1925. Assigned to Western Electric Company, Incorporated.



NUMBER 1,551,624—Circuits for Wave Transmission

CIRCUITS FOR WAVE TRANSMISSION where an electron tube oscillator is coupled with an antenna ground system through a variometer. The variometer may be readily shifted for changing the frequency of the generated oscillations and automatically adjusting the coupling with the antenna system for controlling the production of oscillations.

1,551,661—C. G. Hill, filed May 20, 1922, issued September 1, 1925.

VARIABLE CONDENSER where the plates are arranged in spiral formation in such manner that they may be varied in spaced relationship by rotative movement of a central shaft to vary the capacity of the condenser.

* Received by the Editor, November 15, 1925.

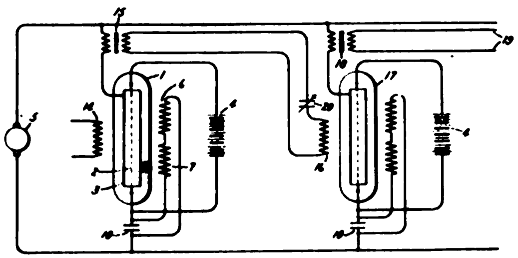
1,551,845—E. L. Popper, filed November 9, 1923, issued September 1, 1925.

CRYSTAL DETECTOR in the form of a cup-shaped unit which may be secured to a panel with a sensitive crystal secured within the cup-shaped unit between two electrodes. The detector is fixed and is intended for permanent mounting in such manner that it will not vary in its adjustment.

1,552,186—R. S. Alcox, filed July 29, 1923, issued September 1, 1925.

CONDENSER where the stator and rotor plates are formed of semi-circular concentric plates and moved to interleave with each other in an axial direction for varying the capacity of the condenser.

1,552,219—G. G. Mercer, filed April 23, 1920, issued September 1, 1925. Assigned to General Electric Company.



NUMBER 1,552,219—Vacuum Tube Circuits

VACUUM TUBE CIRCUITS in which an electron current is controlled by means of a magnetic field. The current which flows between a cathode and anode within an evacuated vessel also passes through a coil which surrounds the electrodes for producing a constant magnetic field in the space between the electrodes. The variable component of the current in the control circuit is by-passed to the cathode and does not affect the constant polarizarion supplied by the magnetic field.

1,552,266, F. C. Bradley, filed February 19, 1925, issued September 1, 1925.

A TUNING UNIT FOR RADIO RECEIVING SETS, including a condenser with a pair of hingedly mounted inductance units which may be rotated by a cam which is controlled from the shaft which carries the rotatable condenser plate. The cam varies the space relationship between the inductances in accordance with the

movement of the plates of the condenser, and in this way the capacity is varied in proportion to the change in the inductances.

1,522,310—C. J. Kayko, filed July 24, 1923, issued September 1, 1925. Assigned to General Electric Company.

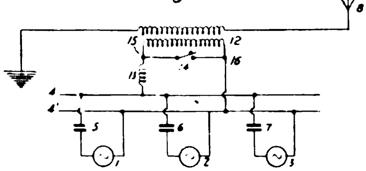
An Electrode for Discharge Tubes which has an operating temperature which is relatively low for a high electron emission. The electron emitting cathode consists largely of nickel containing intimately incorporated therewith a compound of a highly positive metal which is sufficiently refractory to be retained in the cathode at a temperature at which the electron emission is substantial. The cathode obtains a temperature of about 1,200 deg. cent. at 0.1 to 1.0 amperes per square centimeter of surface.

1,552,606—G. Holst and E. Oosterhuis, filed November 4, 1921, issued September 8, 1925. Assigned to Naamlooze Vennootschap Philips' Glohilampen-Farriken, Eindhoven, Netherlands.

An Electron Discharge Device, in which the tube is filled with a gas consisting exclusively of pure argon of a pressure not exceeding one-tenth mm. of mercury. The argon filling is found to have the advantage of low tension on the anode for a maximum strength of signals, and the fact that a grid leak is unnecessary between the grid electrode and the filament when using the gas-filled tube.

1,552,670—G. Belfils, filed August 29, 1921, issued September 8, 1925.

Fig. 2

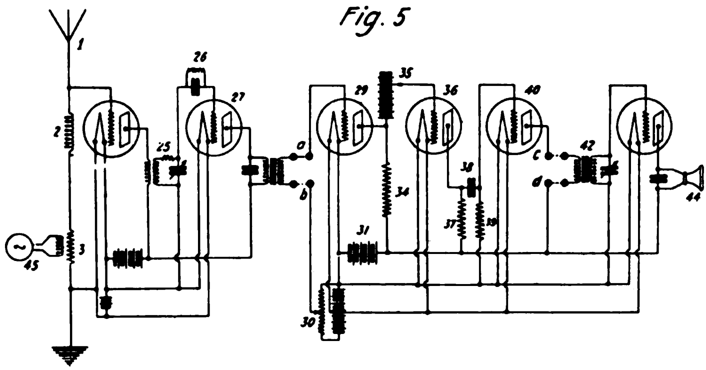


Number 1,552,670—A Radio Transmitting System

A Radio Transmitting System, in which a plurality of alternators are used in connection with the same antenna system. The alternators may be parallel connected to a radiating system and synchronized for operation with respect to each other.

1,552,829—H. J. J. M. De R. De Bellescize, filed August 29, 1921, issued September 8, 1925.

RADIO RECEIVING SYSTEM for the reduction of atmospheric disturbances in the receiving apparatus. A damped receiving circuit is provided where a current-limiting device for receiving the output of the receiving circuit is employed. The current-limiting device comprises a highly damped resonator for eliminating "grinders." A highly resonant circuit for receiving the output of the current-limiting device is provided, and the received signaling energy then impressed upon a responsive device.



NUMBER 1,552,829—Radio Receiving System

1,552,882—F. C. Rowe, filed April 29, 1924, issued September 8, 1925.

TUNING COIL AND CONDENSER MOUNTING FOR RADIO RECEIVERS in which an inductance coil and variable condenser are operated on the same rotatable shaft. The connections for one of the windings of the inductance system are brought out to rings which contact with brushes which are positioned with respect to the rotatable coil in such manner as to make sweeping connection with contacts carried by the rotatable coil for completing connection with an electrical circuit.

1,552,919—Gauthier, Crosby R., filed November 22, 1919, issued September 8, 1925. Assigned to Wired Radio, Inc., New York, New York.

ELECTRICAL COMMUNICATION SYSTEM where broadcasting stations are operated in multiple and modulated over the same interconnecting line wire system. A plurality of broadcasting stations are linked by the same line wire system in this case

and modulated from the same broadcasting studio for simultaneous transmission from the several stations.

1,553,152—R. A. Fessenden, filed November 10, 1920, issued September 8, 1925. Assigned to Submarine Signal Company.

METHOD AND APPARATUS FOR GENERATING ELECTRICAL OSCILLATIONS utilizing two-electrode electron tubes by variation of electron emission of the tubes. A plurality of two-electrode electron tubes are arranged in the form of a Wheatstone bridge circuit and oscillatory variation of the electron emission secured by oscillatory surface heating of the surface of the hot cathodes.

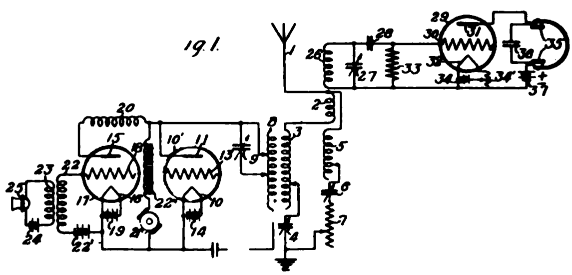
1,553,244—C. F. Jacobs, filed July 10, 1923, issued September 8, 1925.

AN ANTENNA SPREADER for an antenna where the wires are arranged in the form of a cage. The spreader consists of a flat metallic ring having notches at its periphery in which the antenna wires fit and in which they are secured in spaced relationship forming the antenna cage.

1,553,315—L. E. Gould, filed February 28, 1925, issued September 15, 1925.

LOOP FOR RADIORECEPTION in which a circular supporting turn of relatively heavy wire is provided which surrounds a plurality of turns of finer wire which are supported in spaced relationship by insulated members secure upon the heavier turn. The loop is compact in construction and may be inexpensively manufactured on a quantity production basis.

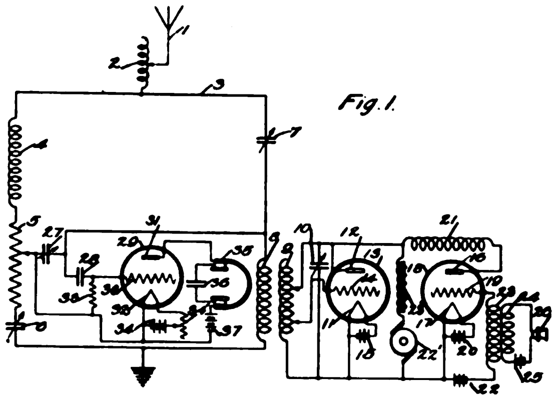
1,553,391—A. Nyman, filed July 15, 1920, issued September 15, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,553,391—Combined Radio Sending and Receiving System

COMBINED RADIO SENDING AND RECEIVING SYSTEM where the effect of the local transmitting energy is neautralized by an auxiliary neutralizing reactance device. The antenna circuit is divided into two branches for connection to the transmitting and receiving apparatus, and undesired currents in the receiving branch are neutralized for preventing interference with the receiving circuit.

1,553,390—A. Nyman, filed July 15, 1920, issued September 15, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,553,390—Combined Radio Sending and Receiving System

COMBINED RADIO SENDING AND RECEIVING SYSTEM where the transmitting and receiving circuits are connected to the same antenna with the receiving apparatus connected at a selected point in the circuit at which there is a zero difference of potential arising from energy derived from the local transmitting station. By selecting a zero potential point interference from side tones is eliminated.

1,553,549—W. H. Priess, filed April 3, 1920, issued September 15, 1925. Assigned to Wireless Specialty Apparatus Company.

CONDENSER which is housed within a relatively heavy casing with a pressure clamp at one side of the casing for subjecting the stack to pressure, while permitting variation in length of the stack for condensers of different capacities.

1,553,454—De Loss K. Martin, filed January 19, 1923, issued September 15, 1925. Assigned to American Telegraph and Telephone Company.

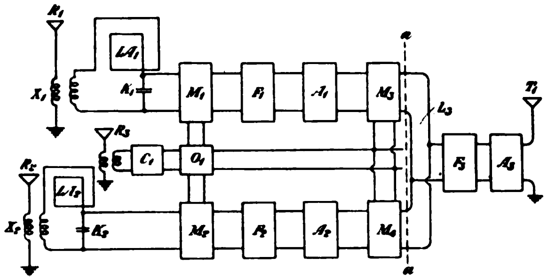
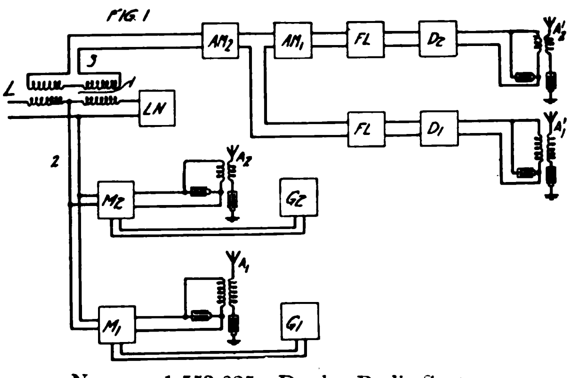


Fig. 4

NUMBER 1,553,454—Radio Repeating System

RADIO REPEATING SYSTEM in which the number of frequencies per channel is limited to two per channel of transmission regardless of the number of repeaters employed. A directly controlled receiving antenna system is arranged to receive from different directions waves having the same frequency or band of frequencies. A frequency translating circuit is provided having means to change the frequency of each of the received waves by the same amount. A directly controlled transmitting system is provided to transmit the waves of changed frequency in different predetermined directions.

1,553,625—J. Mills, filed December 24, 1920, issued September 15, 1925. Assigned to Western Electric Company, Incorporated.



NUMBER 1,553,625—Duplex Radio System

Duplex Radio System where side tone interference at the receiving station due to simultaneous reception of the locally transmitted energy with the desired signaling energy is eliminated by the generation of an auxiliary wave by the local transmitter which opposes current in the receiving circuit induced therein by the local transmitter.

1,553,720—Cyril P. Ryan, filed February 27, 1924, issued September 15, 1925. Assigned to Vickers Limited, Westminster, England.

Radio Signaling and Control system where radio signals are emanated in regularly timed succession at a predetermined acoustic frequency together with the signaling impulses of other acoustic frequencies. A tuned receiving apparatus is arranged to respond only to the timed succession of signals for impressing said signals on a receiving apparatus which has an oscillation period corresponding to the timed succession of impulses for operating a recorder or other apparatus.

1,553,971—S. D. Apostol, filed March 10, 1925, issued September 15, 1925.

Condenser in which a cylindrical roll carries a curved metallic plate on one side thereof which may be moved toward or away from a similarly curved stator plate which is substantially U-shaped for varying the capacity of the condenser.

1,554,231—A. Press, Washington, D. C., filed February 18, 1921, issued September 22, 1925.

Hysteretic Generation of Electromagnetic Waves in a coil antenna system is employed where the antenna is immersed in a mixture of oils. The oils are selected and adjusted to provide a solution having hysteretic and purely dielectric affecting qualities.

1,554,232—A. Press, Washington, D. C., filed January 11, 1922, issued September 22, 1925.

Uniquely Resonant Coil comprising an open-ended helix with a shield extending substantially a one-half stationary wave length along the helix of the frequency of the electromotive force which is applied across the helix. The shield surrounds the helix along a selected portion thereof. The invention may be applied to wave coil systems in which standing waves are set up in an extended coil.

1,554,328—I. M. Brenner, New York, filed November 2, 1923, issued September 22, 1925.

Radio Antenna of the loop type in which the conducting wires forming the loop are carried upon a frame and taps taken from various points in the loop to a selector switch. The selector switch controls a number of effective turns in the loop which may be connected to the radio receiving system.

1,558,436—I. Langmuir, Schenectady, New York, filed October 16, 1913, renewed March 14, 1916, isued October 20, 1925. Asigned to General Electric Company.

Electrical Discharge Apparatus and Process of Preparing and Using the Same in which the tube is evacuated to such a degree that no positive ionization occurs when the impressed voltage is as high as 60 volts and the current over a working range of voltage up to 60 volts varies with the 3/2 power of the impressed voltage. This is the Langmuir patent based on the application which was involved in interference with Arnold on the subject matter of the hard tube.

1,554,598—P. J. Rudy, Centralia, Pennsylvania, filed March 14, 1925, isued September 22, 1925.

Radio Receiver comprising a compact assembly of apparatus which is intended for portable use without the necessity of an antenna ground system. The circuit includes a rectifier sustem, a tuning coil, battery, and telephone receivers, the coil itself serving as a pick-up area for the signaling energy. The rectifier system consists of a plurality of steel, soft iron and lead balls making contact with each other.

1,554,501—L. C. F. Horle, Newark, New Jersey, and N. Heyman, Buffalo, N. Y., filed November 5, 1924, issued September 22, 1925. Assigned to Federal Telephone Manufacturing Corporation.

Shock Proofing Mounting Means for Vacuum Tubes where the electron tubes are carried by a panel which is resiliently suspended on helical springs positioned between the stationary frame of the apparatus and the panel.

1,554,640—V. T. Miller, Kansas City, Mo., filed March 10, 1923, issued September 22, 1925.

Crystal Detector for Radio Receiving Sets where the crystal is secured in a metallic mounting having a plurality of

parallel grooves across the face thereof. The parallel grooves or corrugations have their ends overlapping the edges of the crystal, thereby firmly securing the crystal in the holder.

1,554,713—R. C. Clinker, Inwoods Rugby, England, filed July 10, 1920, issued September 22, 1925. Assigned to General Electric Company.

ELECTRICAL APPARATUS having a rotatable shaft which may be driven through coarse and fine adjustments by moving the control element of the shaft longitudinally with respect to the elements driven by the shaft, so that vernier or micrometer adjustment may be made in the apparatus unit driven by the shaft. The patent shows the invention as applied to a variable condenser.

1,554,720—C. V. Ferguson, Schenectady, New York, filed December 3, 1918, issued September 22, 1925. Assigned to General Electric Company.

ELECTRIC ARCH DEVICE, in which a thermionic discharge tube is provided with an anode consisting of an alloy of alkali metal which is vaporized during the discharge. The alkali vapor permits the operation of an arc-like discharge therein with a voltage of the order of one volt for current of about one ampere.

1,554,795—L. De Forest, New York. Filed May 10, 1915, issued September 22, 1925. Assigned to De Forest Radio Telephone and Telegraph Company.

RADIO SIGNALING SYSTEM, where an electron tube is employed for controlling the output of an arc transmitter. The electron tube has its output circuit connected in shunt with the antenna system and is used as a modulating device for the output of the arc.

1,555,251—W. H. Priess, Belmont, Massachusetts. Filed June 3, 1921, issued September 30, 1925. Assigned to Wireless Specialty Apparatus Company.

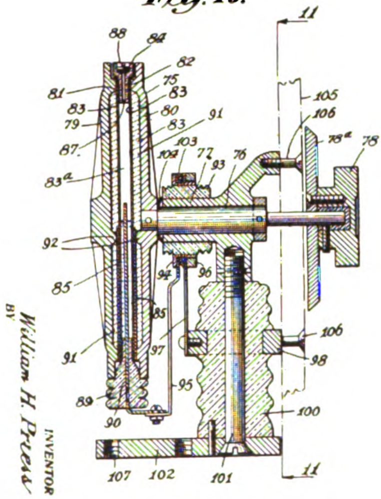
ELECTRICAL CONDENSER design for use in pack sets for the Government where the stack is arranged laterally of a casing and placed under pressure by clamps positioned on opposite sides of the casing.

1,555,252—W. H. Priess, Belmont, Massachusetts, filed June 7, 1921, issued September 30, 1925. Assigned to Wireless Specialty Apparatus Company.

Electrical Condenser for high-potential operation where the plates are built up in stacked formation where a stiff plate is interposed in the stack between the ends thereof and a clamp external to the stack for engaging the plate to brace the plate and the stack. The clamp may also be used to compress the stack from end to end.

1,555,253—W. H. Priess, Belmont, Massachusetts, filed January 6, 1922, issued September 30, 1925. Assigned to Wireless Specialty Apparatus Company.

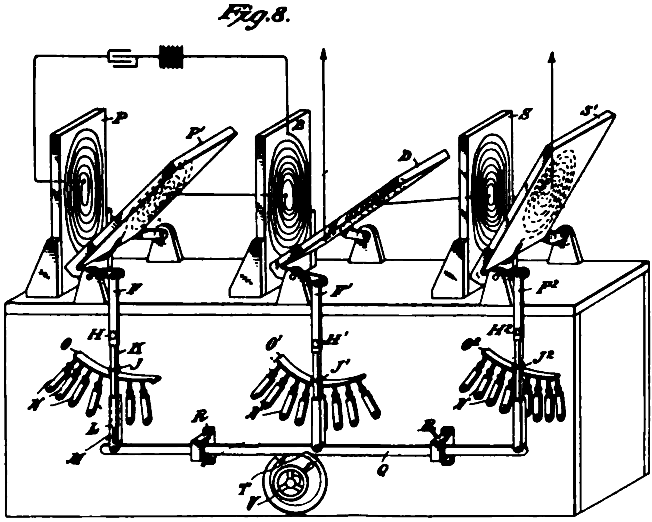
Fig. 10.



Number 1,553,253—Variable Electrical Condenser

Variable Electrical Condenser, where a body of mercury is enclosed within a rotatable reservoir which contains conductive side plates on opposite sides thereof. The reservoir may be moved to different positions to displace the mercury from the reservoir for providing different overlapping relationships between the mercury and the side plates for varying the capacity of the condenser.

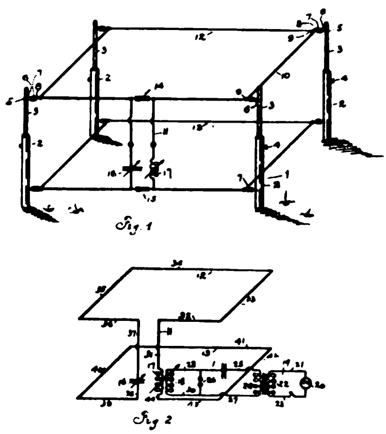
1,555,254—J. A. Proctor, Lexington, Massachusetts. Filed September 4, 1920, issued September 29, 1925. Assigned to Wireless Specialty Apparatus Company.



NUMBER 1,555,254—Electrical Tuning Apparatus

ELECTRICAL TUNING APPARATUS comprising a plurality of separate tuning elements which are connected by a common adjuster and the several elements simultaneously controlled from a single hand actuator. Suitable primary adjustments may be made in each of the tuning elements so that simultaneous effect of all of the tuning elements will be to efficiently control associated electrical circuits.

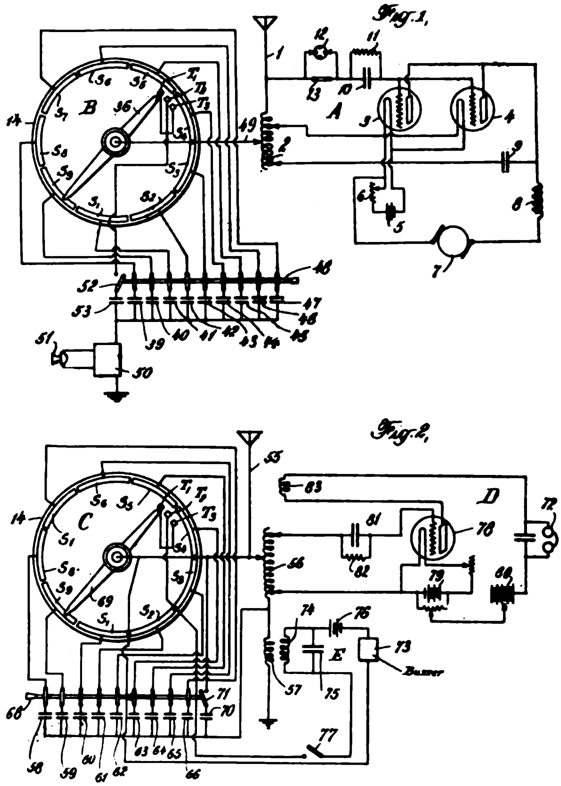
1,555,345—J. A. Willoughby, Washington, D. C., filed June 26, 1919, issued September 30, 1925.



NUMBER 1,555,345—Loop Antenna

Loop Antenna for polarized wave transmission. A plurality of oscillatory loops are provided with substantially closed metallic circuits. A transmitter is arranged to excite the loops in opposition to each other. The loops are spaced substantially parallel and produce by such opposition a field of magnetic disturbance which emanates in a controllable direction.

1,555,633—L. Burch, Belleville, New Jersey. and M. K. Parkhurst, New York City, filed December 9, 1922, issued September 29, 1925.



Number 1,555,633—Secret Signaling System

Secret Signaling System, in which the frequency of the transmitting station is varied over a range synchronously with the variation in frequency of the receiving station. A gyroscope is provided at each terminal and arranged to drive a switch which varies the frequency of the transmitting and receiving circuits simultaneously for maintaining the circuits in synchronism.

1,555,634—S. Cohen, Brooklyn, New York filed November 8, 1924, issued September 29, 1925. Assigned to General Instrument Corporation.

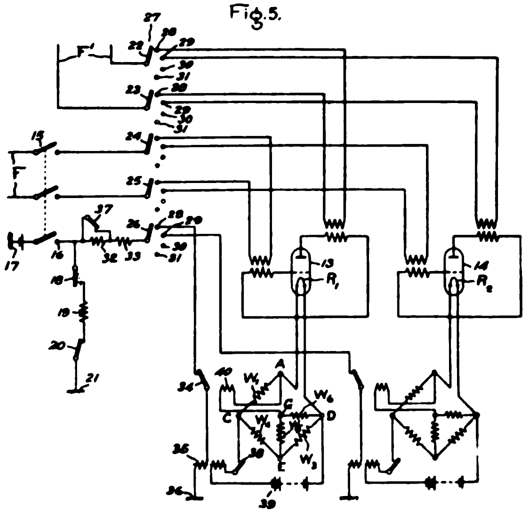
VARIABLE CONDENSER, in which a set of stator plates are mounted in spaced relation within a condenser frame and secured therein by insulators positioned between the frame and the stator plates. The patent illustrates the use of pedestal insulators for supporting the stator plates within the condenser frame with the rotor plates carried by a shaft which is journaled in the frame in such manner that the rotor plates may be inter-leaved with the stator plates. The condenser is designed to have minimum dielectric losses.

1,555,677—C. L. A. M. LeBlanc, Paris, France, filed March 22, 1921, issued September 29, 1925. Assigned to Societe Anonyme pour L'Exploitation des Procedes Maurice-Vickers, Paris.

ELECTRON TUBES for high-power operation where the cathode is in the form of a hollow cylinder and is heated by radiation from a resistance disposed in the interior of the cylinder and raised to incandescence. The grid and anode are also in the form of cylinders co-axial with the cathode. The cathode is heated by thermo-radiation.

1,555,757—G. Respondek, Berlin, Germany, filed November 30, 1925, issued September 29, 1925. Assigned to General Electric Company.

Fig. 5.

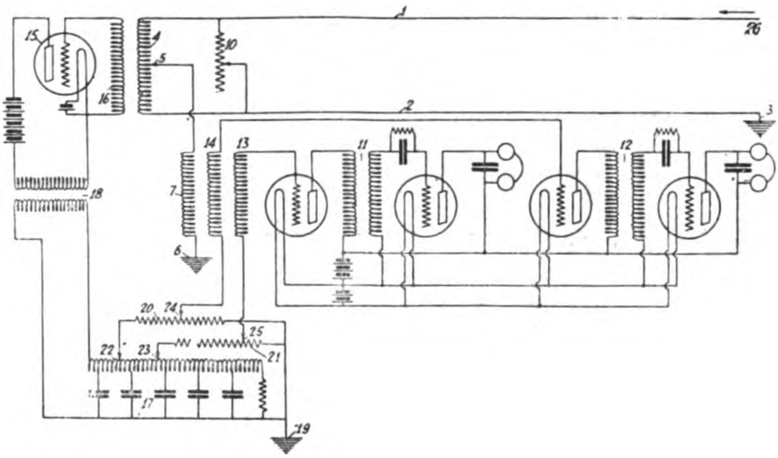


NUMBER 1,555,757—Connection for Vacuum Tubes

Connection for Vacuum Tubes, where a spare tube may be automatically replaced for a tube which has been burned out in the course of normal operation. A Wheatstone bridge circuit is provided where one arm of the Wheatstone bridge is formed by the cathode of the electron tube. When the cathode is destroyed, the bridge is unbalanced, and a relay is operated for connecting the spare tube in the circuit.

1,556,122—A. B. Moulton, New York, filed November 1, 1922, issued October 6, 1925. Assigned to Radio Corporation of America.

Fig. 1

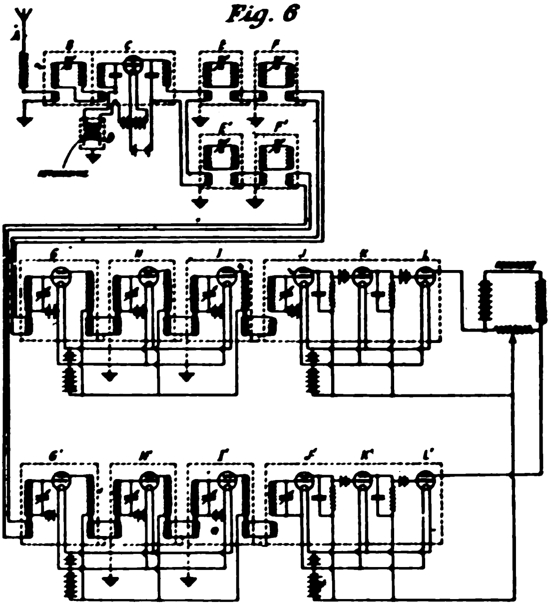


Number 1,556,122—Radio Receiving System

Radio Receiving System, using a Beverage antenna system extending in the general direction of transmission. The antenna system comprises two parallel conductors with a damping circuit connected across a pair of adjoining ends comprising an inductance and a resistance in parallel with a capacity and an inductance in series with the resistance. The circuit is intended for uni-directional reception and for the elimination of interfering signals and atmospheric disturbances. The receiving system is connected to the line wire system at the end opposite the connection of the impedance circuit to the line conductors.

1,556,129—H. J. Round, London, England, filed July 9, 1921, issued October 6, 1925. Assigned to Radio Corporation of America.

Fig. 6



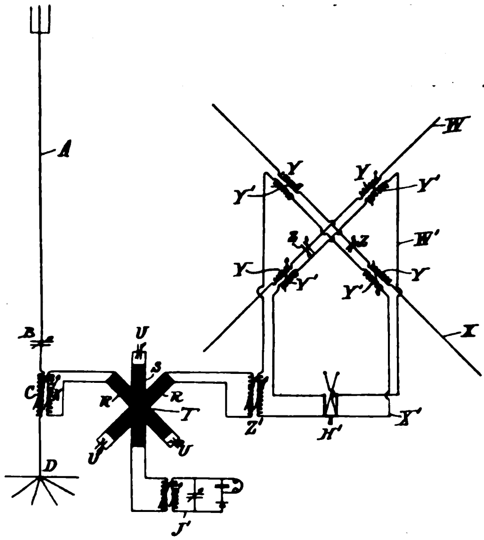
NUMBER 1,556,129—Reception of Radio Signals

RECEPTION OF RADIO SIGNALS by a circuit arrangement designed to be substantially free of atmospheric disturbances. A heterodyne is provided at the receiving circuit to produce beat tones, the currents of which are passed through resonance circuits where undesired disturbances are damped, while the desired signal is carried forward through a system of electron tubes for operating a recorder.

1,556,130—O. Schriever, Berlin, Germany, filed December 27, 1922, issued October 6, 1925. Assigned to Gesellschaft für Drahtlose Telegraphie.

CIRCUIT ARRANGEMENTS FOR RADIO SIGNALING, where transmitting stations are operated at close proximity to each other. A coupling is provided for neutralizing undesired reactive effects between separate transmitting stations which are operated simultaneously.

1,556,137—R. A. Weagant, Douglas Manor, New York, filed February 7, 1919, issued October 6, 1925. Assigned to Radio Corporation of America.



NUMBER 1,556,137—Method and Apparatus for
Radio Signaling

METHOD AND APPARATUS FOR RADIO SIGNALING, whereby static interference may be reduced to a minimum. Two pick-up circuits are provided one of which efficiently receives horizontally propagated signal waves, while the other efficiently receives static impulses as currents substantially in opposite phase to the signaling currents. By this arrangement the static is balanced out while retaining the signal currents.

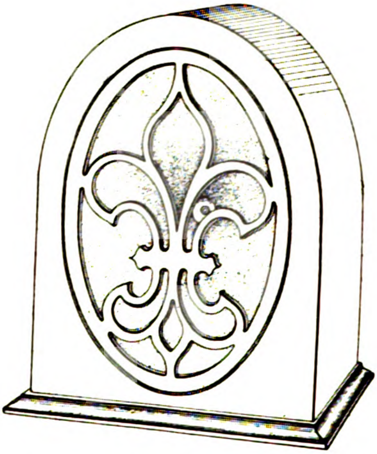
1,556,435—A. S. Gorayeb, New York, filed September 15, 1923, issued October 6, 1925.

PORTABLE ANTENNA, which consists of a casing which may be fitted in the ordinary window sill with an antenna wire carried by a roll within the casing. The antenna wire may be released to extend downwardly from the window frame when the receiving set is placed in operation.

1,556,725—D. H. Shallcross, Claredon, Virginia, filed January 31, 1922, issued October 13, 1925.

SUPPORT FOR RADIO ANTENNAS of the loop type for direction finder work. The support is in the form of a collapsible coil frame having a plurality of members hingedly connected together. In open position the support carries all of the turns of the loop, while in closed position the support may be readily transported from place to place.

Design 68,493—M. C. Rypinski, filed August 6, 1925, issued October 13, 1925. Assigned to Brandes Laboratories, Incorporated, of Newark, New Jersey.



Design 68,493—Radio Reproducer

Radio Reproducer—This patent covers the Brandes cone speaker where an elliptical cone of relatively small size is housed within an ornamental cabinet, shaped to conform with the general contour of the diaphragm.

1,556,633—S. Ruben, New York City, filed September 13, 1924, issued October 13, 1925.

Electrical Control Method for trains and other moving vehicles, where transmitting and receiving apparatus are located upon separate trains which may be approaching each other on the same track. The transmitting apparatus may electrostatically transfer its energy to an overhead line system from the moving train, which energy is in turn transferred again to the other moving train for actuating the receiving apparatus on said train for controlling brakes or other signal when the signal strength has reached a proper degree by reason of the close approach of the trains.

1,556,740 — R. B. Woolverton, Washington, D. C., filed (original) July 25, 1921. Divisional, December 10, 1923. Issued October 13, 1925.

Transmission of Radio Signals Employing Undamped Waves developed by an arc generator, where a tuned oscillatory

circuit is shunted around the arc and an antenna circuit inductively coupled with the tuned oscillatory circuit. A variable inductance is connected in series in the antenna circuit. A solenoid is arranged in a keying circuit and connected so that the inductance of the variable inductance device may be varied for destroying the resonance between the oscillatory and antenna circuits for forming signals in the antenna circuit.

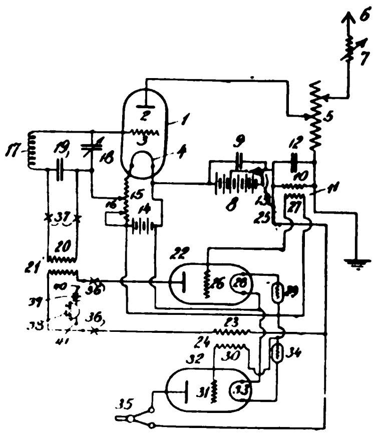
1,556,750—L. B. Bender, Washington, D. C., filed August 29, 1923, issued October 13, 1925.

ELECTRICAL SIGNALING in which the dots and dashes of the Morse code are transmitted and received at different frequencies and combined to actuate a siphon recorder for recording the received signals on a tape.

1,557,049—J. H. Hammond, Jr., Gloucester, Massachusetts, filed May 10, 1918, issued October 13, 1925.

ELECTRICAL ANTENNA for ship use, and particularly submarines, where the antenna is carried in an elongated buoy tube arranged to float on the surface of the water and trail the ship. The antenna wire is carried within the tube and connections established with the apparatus aboard the moving vessel.

1,557,067—L. Kuhn, Berlin-Charlottenburg, Germany, filed August 26, 1921, issued October 13, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,557,067—Combined Transmitting and Receiving Arrangement

COMBINED TRANSMITTING AND RECEIVING ARRANGEMENT where an electron tube is arranged to function both as an oscil-

lator and a detector and is coupled to an antenna system. Switching means are provided whereby the same tube circuits may be employed for reception or transmission.

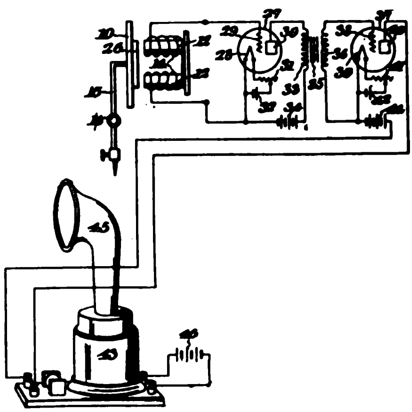
1,557,316—George H. Nobbs, Watertown, Massachusetts, filed December 16, 1924, issued October 13, 1925.

Variable Condenser, in which a selected number of stator plates may be securéd in an electrical circuit in which the condenser is connected. The condenser is of the rotatable plate variety and a switch is mounted adjacent the stator plates for establishing peripheral contact with selected plates in order to include a desired number of plates in the circuit.

1,557,389—E. N. Todd, Crisfield, Maryland, filed March 16, 1925, issued October 13, 1925.

Means for Ascertaining Elevations of Aircrafts comprising a radio transmitting unit which may be dropped from an aircraft and the circuits thereof automatically completed by impact of the apparatus with ground. The aviator may pick up the signals which are radiated from the transmitting apparatus which he has released and by suitable calculation determine his distance above the earth.

1,557,529—E. T. Jones, filed December 3, 1921, issued September 15, 1925.



Number 1,557,529—Electrical Reproducer for Phonographs

Electrical Reproducer for Phonographs where a diaphragm is actuated to vary the magnetic reluctance of a telephone circuit for generating electrical energy in a pair of associated

windings. The energy is amplified and reproduced in accordance with the vibrations of the diaphragm.

1,557,724—W. H. Priess, Belmont, Massachusetts, filed August 2, 1921, issued October 20, 1925. Assigned to Wireless Specialty Apparatus Company.

MACHINE AND METHOD FOR BUILDING ELECTRICAL CONDENSER STACKS by building up alternately a dielectric sheet and a metallic foil sheet and flattening the foil in place on the dielectric sheet by applying a gas under pressure to the said foil. The fixed condensers may be manufactured inexpensively on a quantity production basis.

1,557,725—J. A. Proctor, Lexington, Massachusetts, filed February 1, 1921. Assigned to Wireless Specialty Apparatus Company.

VARIABLE ELECTRICAL CONDENSER where the condenser is housed in a vacuum container, and the moving condenser plates rotated in varying degrees by means of a magnet which is moved exterior of the casing which houses the condenser.

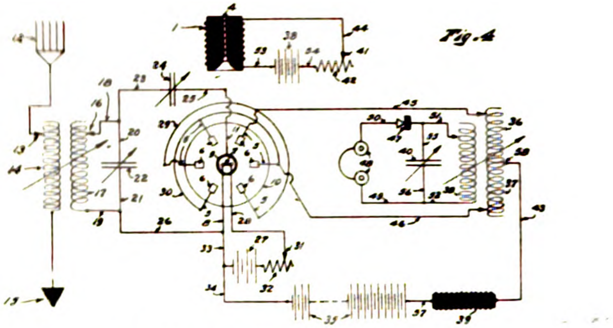
1,558,043—W. H. Priess, Belmont, Massachusetts, filed April 28, 1921, issued October 20, 1925. Assigned to Wireless Specialty Apparatus Company.

ELECTRICAL CONDENSER for high-power operation in which concentric metallic armatures are insulated one from another and embedded in sulphur which provides high insulation resistance with low dielectric loss.

1,558,111—H. E. Metcalf, San Leandro, California, filed March 23, 1925, issued October 20, 1925. Assigned to The Magnavox Company.

VACUUM TUBE in which the grid electrode is formed from a flat plate having a plurality of arms on each edge thereof. The arms are bent in relatively opposite directions to form a trough along each edge of the plate member between the arms, in which trough the filament electrode is stretched. This construction of tube if desirable from the viewpoint of manufacture and assembly for the filament is not centered betwen the electrodes.

1,558,120—F. G. Simpson, Seattle, Washington, filed April 3, 1921, issued October 20, 1925.



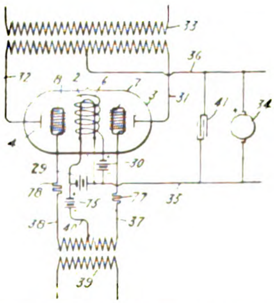
NUMBER 1,558,120—Radio Receiving System

RADIO RECEIVING SYSTEM, in which an alternating current generator consisting of an electron tube system is provided at the receiver and a magnetic field established transverse to the electron stream for varying the velocity of the electron stream in accordance with incoming signaling energy for correspondingly varying the frequency of the alternating current generator and operating a suitable observing circuit.

1,558,144—H. Chireix, Paris, France, filed August 29, 1921, issued October 20, 1925.

ELECTRIC RELAY, comprising an oscillation generator having two oscillating circuits connected thereto and tuned to different frequencies. The generator may be caused to oscillate at either of the frequencies separately. The relay may be used in various circuit arrangements.

1,558,437—I. Langmuir, Schenectady, New York, filed October 29, 1913, renewed February 29, 1924, issued October 20, 1924. Assigned to General Electric Company.



NUMBER 1,558,437—Electrical Discharge Apparatus

ELECTRICAL DISCHARGE APPARATUS in which an auxiliary conductor is provided within a three-electrode electron tube adjacent to the cathode and maintained at a substantially uniform positive potential with respect to the cathode. By this arrangement the effect of space charge in an electron discharge device is reduced, the effect of negatively charged bodies in the proximity of the cathode is eliminated, the discharge current with a given applied voltage is increased and electrons having a relatively uniform velocity are developed.

1,558,535—P. D. Delany, South Orange, New Jerssy, filed February 1, 1922, issued October 27, 1925. Assigned to International Telepost Company, Incorporated.

SECRET RADIO SYSTEM, in which fragmentary parts of the signaling energy which make up telegraphic characters or vocal or instrumental sounds are transmitted at separate frequencies and pieced together at the receiver on a visual recorder for combining the fragmentary parts into an intelligible signal.

1,558,830—Wm. R. Brough, East Orange, New Jersey, filed May 13, 1922, issued October 27, 1925. Assigned to Western Electric Company, Incorporated.

ELECTRON DISCHARGE DEVICE of high power size, where the plate electrodes are cooled by means of a water jacket which is secured by means of a screw collar around the exterior of the tube.

1,558,883—Wm. G. Housekeeper, New York, New York, filed April 20, 1921, issued October 27, 1925. Assigned to Western Electric Company, Incorporated.

A VACUUM TUBE, in which the electrodes are constructed and supported away from the center of the electron tube. A support for the tube electrodes is provided where the electrodes are substantially removed from the center of the tube and supported more nearly adjacent the cylindrical sides of the vessel which houses the tube electrodes.

1,558,961—Wm. R. Bullimore, London, England, filed November 24, 1924, issued October 27, 1925.

MANUFACTURE OF FILAMENT OR CATHODES FOR ELECTRIC LAMPS, THERMIONIC TUBES AND THE LIKE, in which a core of relatively high specific resistance and melting point is coated with a noble metal and an active coating of one or more compounds of the alkaline earth metals.

1,559,116—W. A. Marrison, East Orange, New Jersey, filed October 16, 1924, issued October 27, 1925. Assigned to Western Electric Company, Incorporated.

Wave Generating and Modulating System Employing Quartz Piezo Electric Crystal Circuits in which a plurality of frequencies are generated by piezo electric crystals which frequencies react to produce an audio frequency current under control of the piezo electric crystals.

1,559,193—Maurice W. Stavrum, Robert L. Olson and Wallace H. Berry, Chicago, Illinois, filed August 25, 1924, issued October 27, 1925.

Folding Loop Antenna, in which the frame members of the loop are pivotally mounted on a central clock with slotted end supports at the extremities of the frame members for spacing the turns of the loop with respect to each other. The loop frame may be folded into a small compass.

1,559,280—V. L. Ronci, Brooklyn, New York, filed March 23. 1923, issued October 27, 1925. Assigned to Western Electric Company.

Electron Discharge Device in which the electrodes are supported on threaded metallic members which extend through apertures in a flat insulated disk. The electrodes are secured upon the threaded members by means of wires which are formed into a helix and threaded upon the screw threaded members.

1,559,404—Paul Bunet, Paris, France, filed April 14, 1921, issued October 27, 1925.

High Tension Electrical Condenser in which tubular rod members are provided as the condenser electrodes. The rods making up each side of the condenser are positioned in the form of rings permanently electrically connected together so as to prevent sparking between parts. The distance between successive rods gradually increases throughout the area of the condenser.

1,559,460—S. Ruben, New York, New York, filed June 30, 1920, issued October 27, 1925.

Electron Tube in which a pair of anodes are provided with one of said anodes having an opening therethrough for the passage of electrons from an electron source within the tube. The electron stream is controlled by an external control member which is arranged adjacent the anodes. The tube is intended for high-power transmission.

VOLUME 13 FEBRUARY, 1925 NUMBER 1

PROCEEDINGS
of
The Institute of Radio Engineers

EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY
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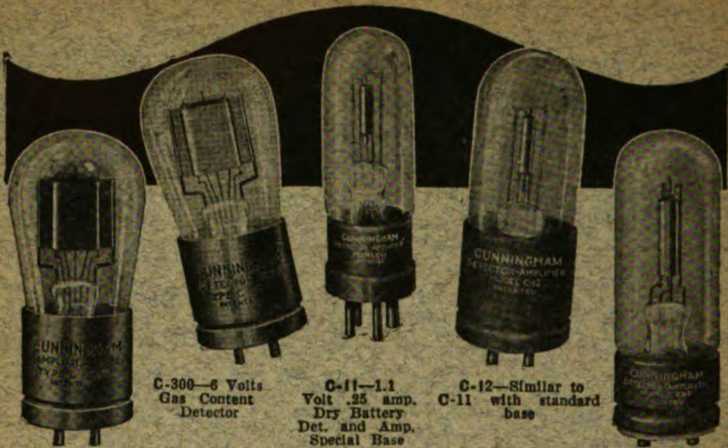

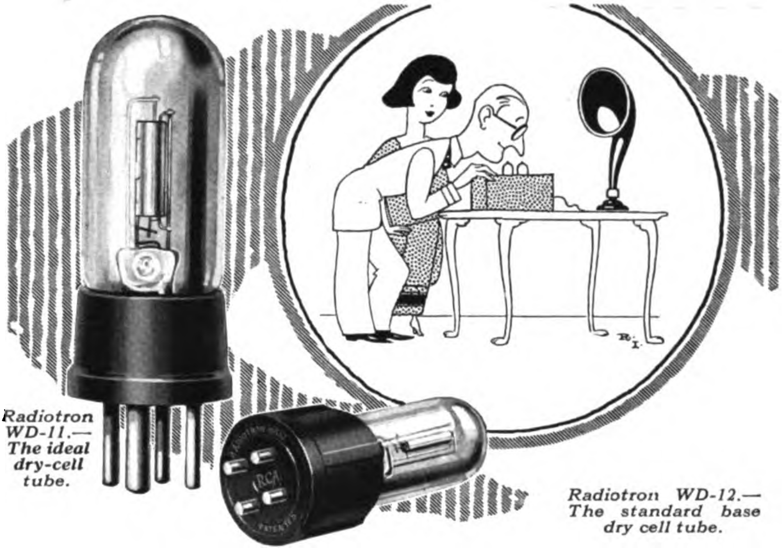

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Radiotron WD-12.—The standard base dry cell tube.

Get a good detector

Radiotrons WD-11 and WD-12 are the same tubes, but with different bases.

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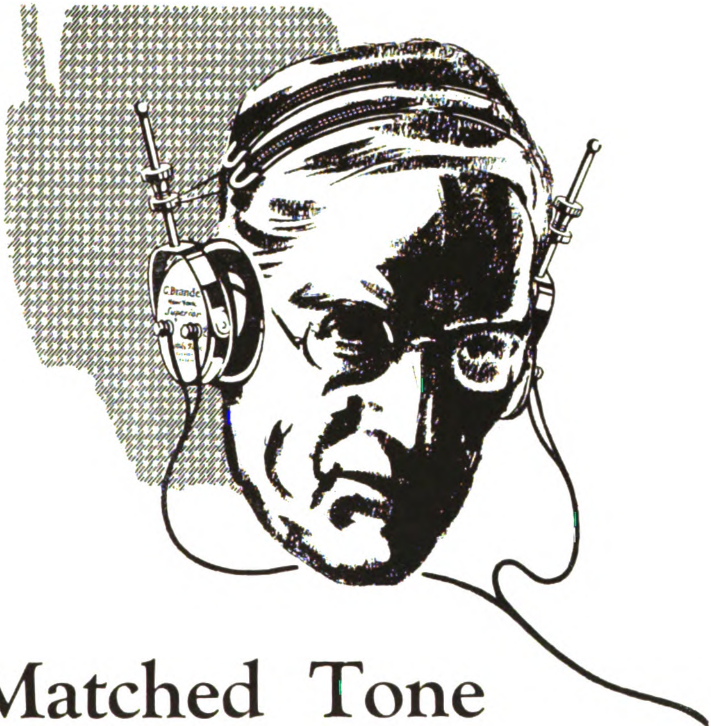
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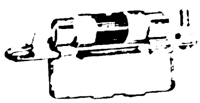

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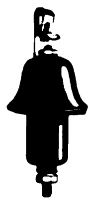

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VOLUME 13 DECEMBER, 1925 NUMBER 6

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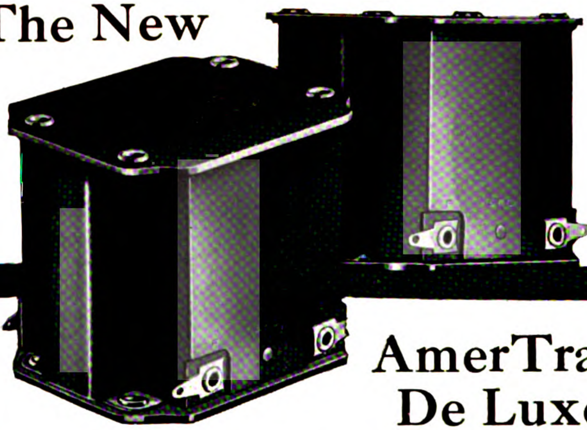
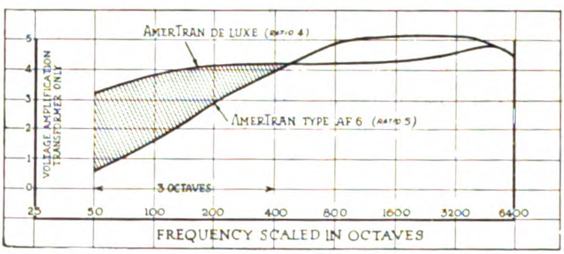
The New



AmerTran De Luxe

THE new De Luxe model AmerTran audio transformer possesses an unusually straight line frequency characteristic extending the range below the lowest note now being broadcast. While the AmerTran AF-6 and AF-7 have, for years, been considered the leaders in audio frequency amplification, this new De Luxe AmerTran shows a gain of about three octaves below that previously obtained.



The AmerTran De Luxe is a transformer of moderate size and weight, enclosed in a strong metal case with mounting holes at both top and bottom so that it may be inverted, affording simplified connections. While the AmerTran De Luxe will improve any set, appreciation of its uniform amplifying qualities can best be realized when operated in conjunction with straight line frequency loudspeakers, such as the best cone and disc types and with a tube in the last stage capable of handling the output. It is for those who are satisfied only with the utmost in quality that this transformer has been developed.

The AmerTran De Luxe is made in two types, one for the first stage and one for the second stage, and plainly marked as such. The chief difference between these two types is that the first stage transformer has approximately 50% greater primary inductance than the second stage transformer, thus more nearly corresponding to the operating impedances of the tubes out of which they work. For this reason it is advisable to purchase and operate these transformers by the pair!

Price, either type, $10.00 at any authorized AmerTran Dealer

AMERICAN TRANSFORMER COMPANY

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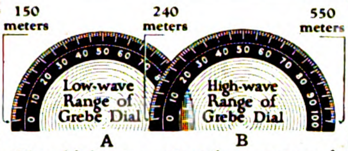
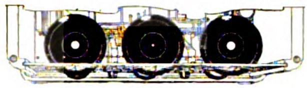
"Transformer builders for over twenty-four years"

AmerTran Models AF-6 and AF-7 now reduced to $5



The high-wave reception range of the Grebe dial (B)—from 550 down to 240 meters—equals the practical tuning range of the usual receiver. The low-wave range of the Grebe dial (A) provides additional reception down to 150 meters.



Flexible Unit Control

One dial operates two or all three; or dials may be adjusted separately, at will.

Grebe "Colortone"

Grebe Binocular Coils
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and
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The Three Outstanding Radio Advances of 1925

LAST year Grebe developed the fieldless *Binocular Coils* and *S-L-F Condenser.* This year even greater strides have been made in radio reception.

With the *Grebe "Colortone"* the quality of sound can be kept absolutely true and clear and is always under the listener's control, independent of the loud speaker's influence. This tone control also makes it possible to reduce, to a great degree, disturbances due to static and to suppress considerably the high-pitch frequencies caused by heterodyne interference of one station with another.

The *Low-Wave Extension Circuits* have widened the reception range so that over 100 low-wave stations not reached by other receivers can be tuned.

Flexible Unit Control gives one, two or three-dial control.

Both in range and quality of reception, the Grebe Synchrophase is far in advance.

Write for literature

A. H. Grebe & Co., Inc., 109 West 57th St., N. Y.

Factory: Richmond Hill, N. Y.

Western Branch: 443 So. San Pedro St., Los Angeles, Cal.

The Grebe Synchrophase

TRADE MARK REG. U.S. PAT. OFF.

This company owns and operates stations WAHG and WBOQ; also low-wave rebroadcasting stations, mobile WGMU and marine WRMU.

All Grebe apparatus is covered by patents granted and pending.

Much toil and great diligence have made the Synchrophase worthy of first rank in radio receivers.

The Synchrophase is also supplied with base for batteries.

For Radio Production Work

Weston has developed special instruments to measure, with the utmost precision and dependability, the minute and peculiar electrical forces encountered in this field. Many of these problems involve the measurement of high frequency currents of very small values, and the Weston line of instruments for radio production and research, though only a few of them are here mentioned, will be found to thorougly cover all needs.

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Model 24

D. C. Switchboard Voltmeter with special scale for direct measurement of resistance. For fast testing, inspection of grid leaks and for work with high resistance apparatus.

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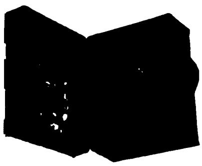
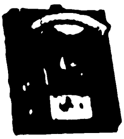


MICROFARADMETER
Model 372

For quick measurement of electrostatic capacities. Essential in production and inspection of condensers or other apparatus which must be held within definite capacity limits.



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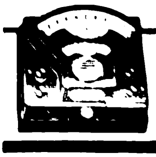

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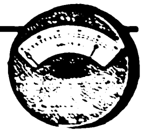
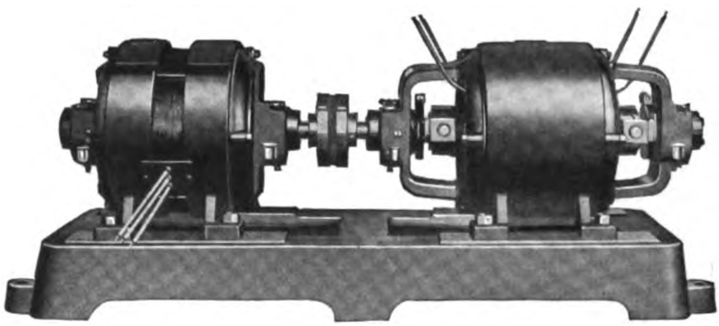


Always specify —
"ESCO" Maximum miles per watt Power supplies for transmission.

"ESCO" has developed a line of over 100 standard 2-bearing Motor-generators for plate or filament. These include D.C., A.C., single phase and polyphase motors.



"ESCO" two and three unit sets have become the accepted standards for transmission. The "ESCO" line consists of over 200 combinations. These are covered by Bulletin 237B.

Our engineers are always willing to cooperate in the development of special sets.

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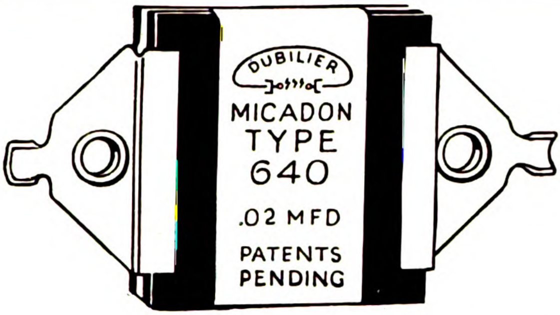

MICADONS—the standard fixed Radio condensers—are shielded by Bakelite

When William Dubilier set out to make a perfect fixed condenser, it was commonly supposed that eddy currents, brush discharges and other hysteresis losses were chiefly responsible for condenser troubles.

Dubilier discovered that compression overcame mechanical movements. This discovery made possible an efficient condenser for broadcasting. It is this same principle that is embodied in the Micadon.

After exhaustive tests, Dubilier found that Bakelite laminated was the best insulating medium obtainable for use in broadcasting condensers and Micadons. Bakelite laminated will not break, does not warp and has high qualities as a dielectric. The development of the Micadon has been an outstanding achievement in which Bakelite has played an important role.

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BAKELITE

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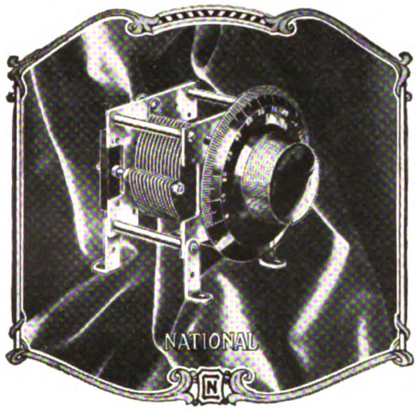

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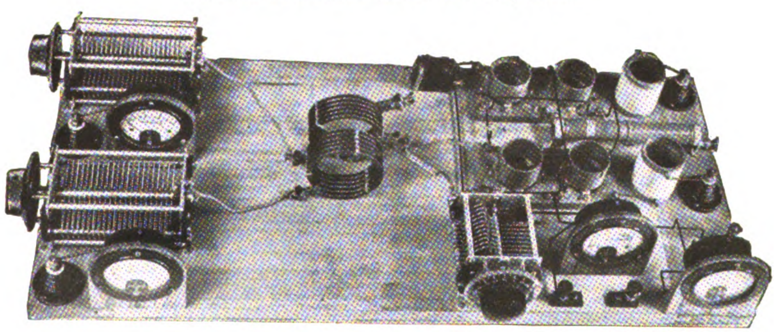

The illustration shows Lieut. F. H. Schnell's short wave Transmitter using National Type D. X. T. transmitting condensers. These condensers are designed for continuous operation at 1500 volts—40 meters

Write for Bulletin 106 I. R.

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